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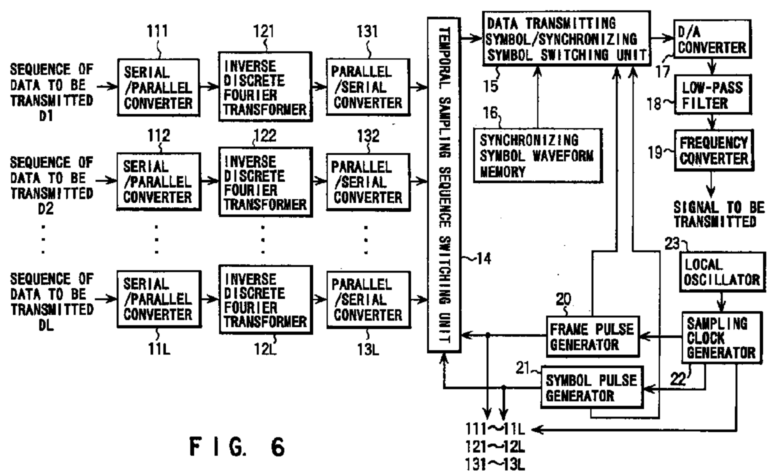
(71) Applicant: **Advanced Digital Television Broadcasting Laboratory**
Tokyo (JP)

(72) Inventors:
- **Saito, Masafumi**
  Tokyo (JP)
- **Ikeda, Tetsuomi**
  Machida-shi, Tokyo (JP)

(74) Representative: **Lewald, Dietrich, Dipl.-Ing. et al**
Lewald . Grape . Schwarzensteiner
Patentanwälte
Rindermarkt 6
80331 München (DE)

(54) MCM system, with a plurality of data streams, each with its own transmission parameters

(57) Data sequences (D1 through DL) correspond to L different parameter sets (effective symbol length, guard interval length, the number of carrier waves) and transformed into parallel data by respective serial/parallel converters (111 through 11L), which are allocated to respective carrier waves for OFDM and subjected to inverse discrete Fourier transform by inverse discrete Fourier transformers (121 through 12L) to produce sampled values for the transmission waveform in the time domain. The sampled values are transformed into serial sequences of sampled values by parallel/serial converters (131 through 13L) and then into a single temporal sampling sequence by a temporal sampling sequence switching unit (14). A frame synchronizing symbol is added to the temporal sampling sequence and then transformed into an analog base band OFDM signal before it is converted up to a transmission signal. The frequency bandwidth of the OFDM signal is made smaller than a predetermined value defined by the bandwidth of the available transmission channel. As a result, an OFDM signal that can be received well regardless of the mode of reception can be transmitted.



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Description

This invention generally relates to a transmission system for digital broadcasting and, more particularly, it relates to a data transmission system using orthogonal frequency division multiplexing (hereinafter referred to as OFDM) and digital modulation/demodulation. The present invention also relates to a transmitter and a receiver adapted to such a system.

The demand for digitized television broadcasting using ground waves has been increasing to improve the quality of television service. The OFDM transmission system appears to be particularly promising for ground wave digital signal transmission because it is robust against the multipath effect (or the ghost effect for television).

The OFDM transmission system is a variation of the multiple carrier modulation system, with which a transmission signal is produced by combining a large number (tens to thousands) of digitally modulated waves (carrier waves 1 through k) as typically illustrated in FIG. 1. Each of the carrier waves may be modulated in a mode selected from a number of different possible modes including QPSK, 16QAM and 64QAM.

The data transmission using the OFDM transmission system is realized by using transmission symbols as illustrated in FIG. 1, each of which constitutes a unit. Each transmission symbol comprises an effective symbol period and a guard interval. The effective symbol period is a signal period essential to data transmission. The guard interval is a redundant signal period designed to reduce the multipath effect by cyclically repeating the signal waveform of the effective symbol period.

If the gap between any two adjacent frequencies is made equal to the reciprocal number of the effective symbol period for OFDM transmission, the nil point of the frequency spectrum of each digitally modulated wave coincides with the center frequency of the adjacent modulated waves as shown in FIG. 2A so that no cross interference occurs between them. As seen from FIG. 2B, the spectrum of an OFDM signal shows a substantially rectangular profile as a whole. If the effective symbol period is ts and the number of carrier waves is K, then the frequency gap between any adjacent carrier waves is equal to 1/ts while the transmission bandwidth is equal to K/ts.

With the OFDM transmission system, a transmission frame is comprised of tens to hundreds of transmission symbols as shown in FIG. 1. FIG. 3 illustrates a typical OFDM transmission frame. The OFDM transmission frame contains frame synchronizing symbols, if necessary, along with data transmitting symbols. If necessary, it may additionally contain service identifying symbols.

FIG. 4 illustrates the concept of a transmitter A and a receiver B adapted to the OFDM transmission system.

The transmitter A divides a binary data to be transmitted into data blocks, each of which has a predetermined number of bits and is converted into a complex number prior to transmission. Serial/parallel converter A1 allocates different complex numbers Ci (i = 1 to N) to the carrier wave frequencies on a one by one basis and inverse discrete Fourier transform circuit A2 carries out an operation of inverse discrete Fourier transform to the time domain. As a result, sampled data are produced for a time base waveform so that a base band analog signal having a temporally continuous waveform is obtained from the sampled data and processed for frequency conversion by frequency converter A3 before it is transmitted.

The number of sampled values produced on a time base by inverse discrete Fourier transform is typically 2^n for each effective symbol period (n being a positive integer). Thus, if r_G is defined as r_G = (guard interval length)/(effective symbol length), then $2^n \cdot (1+r_G)$ samples are produced for each transmission symbol. The length of each transmission symbol is usually equal to the time interval of sampling points multiplied by an integer.

On the part of the receiver B, frequency converter B1 processes the received signal for frequency conversion to obtain a base band signal waveform, which is sampled at a sampling rate same as that of the transmitter. Discrete Fourier transform circuit B2 processes the sampled data to carry out an operation of discrete Fourier transform to the frequency domain and obtains by calculation the phase and the amplitude of each of the carrier wave frequency components to determine the value of each of the received data before they are converted into serial data by parallel/serial converter B3 and produced as data output.

While television signals are received either in the fixed mode or in the mobile mode (including the portable reception mode), a good reception is essential regardless of the mode of reception. With any known OFDM system, the effective symbol length, the guard interval length and the number of carrier waves of data transmitting symbols are determined mainly on the basis of either the fixed reception mode or the mobile reception mode. If the effective symbol length, the guard interval length and the number of carrier waves of data transmitting symbols are based mainly on, for example, the fixed reception mode, not the mobile reception mode, the influence of fading will be serious.

As pointed out above, with any known OFDM transmission system, the effective symbol length, the guard interval length and the number of carrier waves of data transmitting symbols are determined on the basis of the most popular reception mode because they cannot be selected so as to adapt themselves to more than one different modes.

It is, therefore, the object of the present invention to provide an OFDM transmission system that ensures a good signal reception regardless of the selected reception mode and a transmitter and a receiver adapted to such a system.

According to the invention, the above object is achieved by providing an OFDM transmission system for transmitting data by means of OFDM and digital modulation/demodulation, characterized in that, if the time interval of OFDM sam-

pling points is T, the effective symbol length NiT (Ni being a positive integer), the guard interval length MiT (Mi being zero or a positive integer) and the number of carrier waves Ki (Ki being a positive integer) of the i-th data transmission symbol in an OFDM transmission frame can take a plurality of respectively different values that can be arbitrarily selected, provided that Ki/NiT is kept smaller than a constant value W (W being a positive real number) determined by the bandwidth of the transmission channel.

In other words, with the OFDM transmission system according to the invention, two or more than two values are used for the effective symbol length and also for the guard interval length of a data transmission symbol and the symbol length is made equal to the sampling period, which is a basic unit for OFDM digital signal processing, multiplied by an integer. Additionally, the frequency bandwidth of OFDM transmission signal is made smaller than a constant value determined by the bandwidth of the transmission channel.

As a result, no cross interference appears if a plurality of data transmitting symbols having respective effective symbol lengths and guard interval lengths that are different from each other are multiplexed in a single transmission channel. Thus, the OFDM transmission system according to the invention can meet different conditions for data transmission in a single transmission channel without reducing the efficiency of the use of frequencies and entailing any cross interference among carrier waves.

This invention can be more fully understood from the following detailed description when taken in conjunction with the accompanying drawings, in which:

FIG. 1 is a graph schematically showing signal transmission waveforms and a transmission symbol used for the OFDM transmission system;
FIGS. 2A and 2B are graphs schematically showing the frequency spectrum of the OFDM transmission system;
FIG. 3 is a graph schematically illustrating the configuration of a transmission frame of the OFDM transmission system;
FIG. 4 is a schematic block circuit diagram of a transmitter and a receiver adapted to a known OFDM transmission system;
FIG. 5 is a graph showing the frequency spectrum of an OFDM transmission system having a frequency block for the fixed reception mode and a frequency block for the mobile reception mode in a transmission channel;
FIG. 6 is a schematic block circuit diagram of an embodiment of a transmitter adapted to the OFDM transmission system of the invention;
FIG. 7 is a schematic block circuit diagram of an embodiment of a receiver adapted to the OFDM transmission system of the invention;
FIG. 8 is a graph schematically illustrating the configuration of a transmission frame of the OFDM transmission system according to the invention;
FIG. 9 is a graph schematically illustrating the relationship between the effective symbol length of an OFDM transmission symbol and the average transmission power level that can be used for another embodiment of the invention;
FIGS. 10A and 10B respectively show schematic block diagrams of an embodiment of a transmitter and that of a receiver according to the invention and designed to change the carrier wave frequencies at predetermined periods and a predetermined frequency gap;
FIGS. 11A and 11B show two alternative arrangements of carrier waves for shifting the frequency of each of the carrier waves within a base band in order to change the frequency at predetermined periods, of which FIG. 11A is designed for symbols for mobile reception whereas FIG. 11B is designed for symbols for fixed reception; and
FIG. 12 is a schematic view of a transmission frame comprising one or more data transmitting symbols for mobile and fixed receivers, where symbols for mobile reception are arranged at predetermined time periods.

The idea underlying the present invention will firstly be described. OFDM transmission symbols adapted to fixed reception and those adapted to mobile reception may be transmitted through a single transmission channel by dividing each OFDM signal into two frequency blocks on a frequency base, which frequency blocks are separated by a guard band in order to prevent interference from taking place among carrier waves, and selecting different values for the symbol length in these frequency blocks, which are respectively used for the fixed and mobile reception modes as shown in FIG. 5.

However, with the above method of dividing an OFDM signal into a plurality of blocks, the carrier waves which belong to different frequency blocks can not have an orthogonal relationship because the effective symbol length and the carrier wave frequency gap are differentiated from a frequency block to another, therefore, a guard band has to be provided between any adjacent frequency blocks at the cost of reducing the efficiency of the use of frequencies and the transmission bit rate of a transmission channel.

According to the invention, data transmitting symbols adapted to fixed reception and those adapted to mobile reception can be transmitted through a single transmission channel without providing one or more than one guard bands to prevent cross interference from occurring among carrier waves and hence without reducing the efficiency of

the use of frequencies.

With OFDM transmission systems, an FFT window having the same length as that of the effective symbol period is provided in each data transmitting symbol period and 2^n sampling points are subjected to an operation of discrete Fourier transform to the frequency domain in the demodulator.

The FFT window is arranged usually at the rear end of each transmission symbol. Note that no ghost can get into the FFT window in the demodulator from the adjacent symbol if the multipath delay time (or the ghost signal delay time for television) is shorter than the guard interval length. Therefore, the degradation due to the multipath phenomenon can be made far less serious than the degradation in a single carrier arrangement. Thus, with OFDM transmission systems, the influence of a ghost having a long delay time can be prevented by selecting a long guard interval to make the system substantially unaffected by the multipath phenomenon.

Now, an OFDM symbol length and a guard interval length adapted to fixed reception and those adapted to mobile reception will be discussed below.

Generally, the influence of multipath is one of the most important technological problems that have to be dealt with to achieve a good fixed reception for the OFDM system. As pointed out above, the use of a long guard interval is a useful technique for the prevention of the influence of ghost signals.

However, since the guard interval adversely affects the transmission capacity (bit rate) of a symbol in a manner as described above and the use of a long guard interval reduces the bit rate of a symbol having a given length, the effective symbol length has to be made proportional to the guard interval length and, therefore, a symbol having a long length has to be used to maintain a desired level of bit rate.

For the mobile reception mode, on the other hand, the characteristics of the transmission channel can change with time due to the fading phenomenon and, therefore, the use of a long OFDM symbol can result in an unnegligible change in the characteristics of the transmission channel within the time required for the transmission of a single symbol and hence a large bit error rate appears if the OFDM symbol length is too long. In other words, a long guard interval and hence a long symbol length operate disadvantageously for the fading phenomenon that can be observed in mobile reception. It may be safe to say that the portable reception mode is a combination of the fixed and mobile reception modes.

As discussed above, for the OFDM transmission system, the optimal values of the guard interval length, the effective symbol length and other transmission parameters may vary depending on the mode of reception. Therefore, a single set of values probably cannot optimize the reception in both the fixed and mobile reception modes. Thus, the OFDM transmission system according to the invention will be particularly useful when transmitting data through a single transmission channel for both the fixed and mobile reception modes, while using long symbols in the fixed reception mode.

Now, an embodiment of a transmitter and that of a receiver adapted to the OFDM transmission system according to the invention will be described in detail by referring to FIGS. 6 and 7.

FIG. 6 is a schematic block circuit diagram of an embodiment of a transmitter adapted to the OFDM transmission system of the invention. The transmitter comprises serial/parallel converters 111 through 11L, inverse discrete Fourier transformers 121 through 12L, parallel/serial converters 131 through 13L, a temporal sampling sequence switching unit 14, a data transmitting symbol/synchronizing symbol switching unit 15, a synchronizing symbol waveform memory 16, a D/A converter 17, a low-pass filter 18, a frequency converter 19, a frame pulse generator 20, a symbol pulse generator 21, a sampling clock generator 22 and a local oscillator 23.

A total of L sequential data D1 through DL to be transmitted are applied respectively to the L serial/parallel converters 111 through 11L. A set of L parameters (effective symbol length, guard interval length, number of carrier waves) are provided to correspond to the L sequential data D1 through DL to be transmitted.

Said serial/parallel converters 111 through 11L convert respective serial data into parallel data, which are allocated respectively to the carrier waves of the OFDM transmission system. The inverse discrete Fourier transformers 121 through 12L determines the phases and the amplitudes of the respective carrier waves on the basis of the data allocated for transmission. The phase and the amplitude of each of the carrier waves are treated as a complex number in the frequency domain and subjected to an operation of inverse discrete Fourier transform and the sampled values of the transmission waveform obtained in the time domain are produced as outputs. Then, the parallel/serial converters 131 through 13L convert the temporally sampled values produced in parallel into a sequence of serially sampled values for each symbol.

On the other hand, the sampling clock generator 22 generates a sampling clock on the basis of the original oscillation frequency signal produced by the local oscillator 23. The frame pulse generator 20 and the symbol pulse generator 21 respectively generate a frame pulse and a symbol pulse from the sampling clock. The sampling clock, the frame pulse and the symbol pulse are fed to the components of the transmitter for timing purposes.

The temporal sampling sequence switching unit 14 selectively switches the L sequences of temporal samples to transform them into a single sequence of temporal samples by using the frame pulse and the symbol pulse. The synchronizing symbol waveform memory 16 produces the sampled values of the frame synchronizing symbol waveform. The data transmitting symbol/synchronizing symbol switching unit 15 switches the sequence of temporally sampled values of the data transmitting symbol produced by the temporal sampling sequence switching unit 14 and the sequence

of sampled values of the waveform of the frame synchronizing symbol produced by the synchronizing symbol waveform memory 16 to transform them into a sequence of temporally sampled values of the base band OFDM signal.

The D/A converter 17 converts the sequence of temporally sampled values into an analog signal and the low-pass filter 18 eliminates the high frequency components of the analog signal to produce an analog base band OFDM signal. The frequency converter 19 converts the frequency of the base band OFDM signal up to an intermediate frequency or a radio frequency and produces a signal to be transmitted.

FIG. 7 is a schematic block circuit diagram of an embodiment of a receiver adapted to the OFDM transmission system of the invention. It comprises a band-pass filter 31, a frequency converter 32, a synchronizing symbol waveform memory 33, a synchronizing symbol position detector 34, an oscillation frequency control signal generator 35, a local oscillator 36, a sampling clock generator 37, a frame pulse generator 38, a symbol pulse generator 39, an A/D converter 40, serial/parallel converters 411 through 41L, discrete Fourier transformers 421 through 42L and demodulation-parallel/serial converters 431 through 43L.

In the receiver having the configuration described above, the band-pass filter 31 eliminates the out-of-band components and the frequency converter 32 converts the intermediate frequency or the radio frequency of the OFDM signal down to a base band. The A/D converter 40 transforms the base band OFDM signal into a sequence of sampled digital values, which are respectively fed to the serial/parallel converter 411 through 41L and also to the synchronizing symbol position detector 34.

The synchronizing symbol position detector 34 detects the position of the front end of the frame by calculating the correlated values of the sequence of sampled values of the base band OFDM signal and the sequence of sampled values of the synchronizing symbol waveform stored in the synchronizing symbol waveform memory 33. It also determines the position for switching the transmitting symbols and the position of the FFT window.

The oscillation frequency control signal generator 35 generates a signal for controlling the oscillation frequency of the local oscillator 36 on the basis of the frame period detected by the synchronizing symbol position detector 34. A method of controlling the local oscillation frequency by means of a frame period is described in Japanese Patent Application No. 6-138386 "Clock frequency automatic control method and transmitter and receiver using the same".

The sampling clock generator 37 generates a sampling clock on the basis of the original oscillation frequency signal produced by the local oscillator 36. The frame pulse generator 38 and the symbol pulse generator 39 respectively generate a frame pulse and a symbol pulse on the basis of the data on the position of the front end of the frame produced by the synchronizing symbol position detector 34 and the sampling clock. The sampling clock, the frame pulse and the symbol pulse are respectively fed to the related components of the receiver and used to generate various timing signals.

The serial/parallel converter 411 through 41L transform the sequence of sampled base band values into parallel data, which are fed then to the discrete Fourier transformers 421 through 42L. The discrete Fourier transformers 421 through 42L transform the sampled values in the time domain into spectra for the respective carrier wave frequencies. The demodulation-parallel/serial converters 431 through 43L estimate the phases and the amplitudes of the carrier waves from the respective frequency component, determines the values of the received data on the basis of the phases and the amplitudes and transform them into sequences of serial received data D1 through DL, which are then produced by the converters as respective outputs. The L sequences of received data D1 through DL corresponds to the L parameter sets.

In the transmission system having the configuration described above, the inverse discrete Fourier transformer 12i (i being an integer between 1 and L) and the discrete Fourier transformer 42i (i being an integer between 1 and L) arbitrarily select Ni, Mi and Ki provided that Ki/NiT is kept smaller than a constant value W (W being a positive real number) determined by the bandwidth of the transmission channel, where T is the time interval of sampling clocks, NiT is the effective symbol length (Ni being a positive integer), MiT is the guard interval length (Mi being zero or a positive integer) and Ki is the number of carrier waves (Ki being a positive integer).

The temporal sampling sequence switching unit 14 switches the data transmitting symbols in such an order that the data transmitting symbols having an identical effective symbol length and a guard interval length are continuously arranged on the time base and the number of switching points where two adjacent data transmitting symbols having at least mutually different effective symbol lengths or mutually different guard interval lengths are located is minimized.

While there may be a number of different orders according to which symbols corresponding to data sequences D1 through DL are transmitted, data transmitting symbols corresponding to a sequence of data (a set of parameters) are to be most basically transmitted in an continuous order on the time base. Then, the number of switching points where two adjacent data transmitting symbols having respective sets of parameters that are different from each other are located is minimized. FIG. 8 shows a typical arrangement of data transmitting symbols that meets the above requirements.

Assuming L = 2 and that sequence D1 of transmission data is for the fixed reception mode and sequence D2 of transmission data is for the mobile reception mode, a good data reception can be realized in either mode by selecting respective sets of parameters for fixed reception and mobile reception for the inverse discrete Fourier transformers 121 and 122.

Thus, with the OFDM transmission system according to the invention, any cross interference can be prevented from appearing between two adjacent carrier waves without using a guard band so that various different requirements of transmission can be met within a single transmission channel without reducing the efficiency of the use of frequencies. Specifically, OFDM data transmitting symbols good for fixed reception and those adapted to mobile reception can be transmitted through a single transmission channel without reducing the efficiency of the use of frequencies.

While L inverse discrete Fourier transformers 121 through 12L are used for L different parameter sets in the arrangement of FIG. 6, a single inverse discrete Fourier transformer may cover L different symbol lengths if it is adapted to the use of a plurality of FFT points.

While the technique of modulation to be used for the OFDM carrier waves may be selected depending on the phase and the amplitude assigned to each carrier wave in the form of a complex number in the frequency domain, different techniques of modulation may be respectively used for sequences D1 through DL of transmission data, typically including techniques such as DQPSK, 16QAM and 64QAM.

Similarly, while L discrete Fourier transformers 421 through 42L are used for L different parameter sets in the arrangement of FIG. 7, a single discrete Fourier transformer may cover L different symbol lengths if it is adapted to the use of a plurality of FFT points.

Of the L data sequences D1 through DL, the sequence D1, for example, and the corresponding transmission symbols may be used to transmit data on the effective symbol lengths, the guard interval lengths, the number of carrier waves and the modulation techniques selected for the carrier waves for the remaining data sequences D2 through DL from the transmitter to the receiver.

Generally speaking, if the effective symbol length NaT (Na being a positive integer), the guard interval length MaT (Ma being zero or a positive integer) and the number of carrier waves Ka (Ka being a positive integer) of a specific data transmitting symbol in an OFDM transmission frame are known by the receiver along with the modulation techniques selected for the carrier waves of the specific symbol, the parameter sets for the data transmitting symbols can be modified by transmitting at least part of the data on the effective symbol lengths, the guard interval lengths, the number of carrier waves and the modulation techniques selected for the respective carrier waves of all the data transmitting symbols other than said specific data transmitting symbol in the frame from the transmitter to the receiver by means of said specific symbol.

If the average transmission power required for the i-th data transmitting symbol can be Pi in the OFDM frame of the above embodiment, Pi can be determined as a function of Ni that defines the effective symbol length in the frame so that Pi and Ni provide a one-to-one correspondence. If, additionally, there are L different possible values of Ni, there will also be L different possible values of Pi so that the average transmission power Pi may vary depending on the effective symbol length NiT of each data transmitting symbol.

With such an arrangement, different service areas may be provided for fixed reception and for mobile reception by selecting different values of the average transmission powers for fixed reception and for mobile reception.

If, in the above embodiment, the value of Ni is selected from A1, A2, ..., AL and Amax is the largest value of A1, A2, ..., AL, all the numbers Al, A2, ..., AL may be so selected as to be divisors of Amax that can exactly divide the latter. In other words, if the effective symbol length NiT is selected from A1T, A2T, ..., ALT and AmaxT is the largest value of AlT, A2T, ..., ALT, they are divisors of AmaxT that can exactly divide the latter.

Then, all the data transmitting symbols can commonly share part of the carrier waves. Therefore, data on carrier phase necessary for coherent demodulation or control data can be transmitted by means of such commonly shared carrier waves.

On the other hand, if Mi that defines the guard interval length can take only a single value, that is, if there are a plurality of values that the effective symbol length NiT can take and the guard interval length MiT can take only a single value, then the layer for fixed reception and that for mobile reception in each transmission frame will perform exactly in a same manner against inter-symbol interference due to multipath.

Finally, the frequencies of the carrier waves for the data transmitting symbols in an OFDM transmission frame may be shifted with a predetermined period and a predetermined frequency interval. More specifically, the frequencies of the carrier waves of data transmission symbols having a relatively few number of carrier waves (or symbols for mobile reception) may be shifted by the frequency interval of the carrier waves of other data transmission symbols having a relatively large number of carrier waves (or symbols for fixed reception) multiplied by an integer.

With such an arrangement, data on carrier phase necessary for coherent demodulation of symbols for fixed reception or data for equalization can be transmitted by means of the carrier waves of symbols for mobile reception.

More specifically, the frequencies may be shifted in a radio frequency band or in a base band. FIGS. 10A and 10B are circuit configurations adapted to the former, whereas FIGS. 11A and 11B illustrates possible arrangements of carrier waves adapted to the latter. The components common to those of FIGS. 6 and 7 are respectively denoted by the same reference symbols.

(1) Frequency Shift in a Radio Frequency Band

FIG. 10A shows a circuit configuration of a transmitter designed for a frequency shift in a radio frequency band. Variable frequency local oscillator 24 shifts the oscillation frequency according to the transmission symbol by means of the frame pulse and the symbol pulse fed respectively from the frame pulse generator 20 and the symbol pulse generator 21 shown in FIG. 6. The frequency converter 19 shown in FIG. 6 can be driven by the shifted oscillation frequency to generate, for each data transmitting symbol, signals having frequencies shifted.

FIG. 10B shows a circuit configuration of a receiver designed for a frequency shift in a radio frequency band. Variable frequency local oscillator 44 shifts the oscillation frequency according to the transmission symbol by means of the frame pulse and the symbol pulse fed respectively from the frame pulse generator 38 and the symbol pulse generator 39 shown in FIG. 7. The frequency converter 32 shown in FIG. 7 can be driven to convert the OFDM signal having an intermediate or radio frequency down to a base band.

(2) Frequency Shift in a Base Band

FIG. 11A shows an arrangement of carrier wave frequencies of symbols for mobile reception. FIG. 11B shows an arrangement of carrier wave frequencies of symbols for fixed reception. In the example shown in FIGS. 11A and 11B, m = 10 and n = 40, where m is the number of carrier wave frequencies provided for the mobile-reception symbols, and n is the number of carrier wave frequencies provided for the fixed-reception symbols.

Numerals 1 to n are assigned to the carrier-wave frequencies at the input of the inverse discrete Fourier transformer provided in the transmitter. The numerals will be referred to as "frequency slot numbers." Further, numerals 1, 2, 3, ... are assigned to the symbols for data transmission. In the symbol 1 for mobile reception, data is set to the inverse discrete Fourier transformer, in every (n/m)th slot, starting with the slot 1; in the symbol 2, data is set in every (n/m)th slot, starting with the slot 2, and so forth. The inverse discrete Fourier transformer which has n points converts the data to the time domain. A signal having frequencies shifted along the time axis as shown in FIG. 11A is thereby generated.

In the symbols for fixed reception, data are set at all n points of the inverse discrete Fourier transformer, as is illustrated in FIG. 11B. The transformer performs inverse discrete Fourier transform on the data.

On the demodulation side, too, a discrete Fourier transformer having n points is used. After the discrete Fourier transform has been performed, a symbol for mobile reception is demodulated by selecting only the frequency slots which are used in the symbol.

In the embodiment described above, one or more data transmitting symbols for mobile reception which have relatively short effective length and guard interval length may be transmitted at regular time intervals within a frame, as is illustrated in FIG. 12. The embodiment can thereby realize a time interleaving effect against fading in mobile reception. This reduces burst errors and also the memory capacity required for interleaving, than in the case where symbols for mobile reception are transmitted continuously.

The above embodiment may be used for ATM telecommunications where the amount of data to be transmitted can vary as a function of time and also for transmission of data encoded in the form of variable length codes, by modifying the transmission parameters including the effective symbol length, the guard interval length, the number of carrier waves and the modulation techniques for each carrier wave by means of a specific data transmitting symbol to modify the amount of data to be transmitted on a frame basis.

The present invention is not limited to the above embodiment, which may be subjected to various changes and modifications.

Thus, the invention provides an OFDM transmission system that ensures a good reception regardless of the mode of reception and a transmitter and a receiver adapted to such a system.

Claims

1. An orthogonal frequency division multiplexing (hereinafter referred to as OFDM) transmission system for transmitting data by means of OFDM and digital modulation/demodulation, characterized in that, if the time interval of OFDM sampling points is denoted as T, the effective symbol length NiT (Ni being a positive integer), the guard interval length MiT (Mi being zero or a positive integer) and the number of carrier waves Ki (Ki being a positive integer) of the i-th data transmission symbol in an OFDM transmission frame, Ni, Mi and Ki can take a plurality of respective different values that can be arbitrarily selected, provided that Ki/NiT is kept smaller than a constant value W (W being a positive real number) determined by the bandwidth of the transmission channel.

2. An orthogonal frequency division multiplexing transmission system according to claim 1, characterized in that, if the effective symbol length NaT (Na being a positive integer), the guard interval length MaT (Ma being zero or a positive integer) and the number of carrier waves Ka (Ka being a positive integer) of a specific data transmitting symbol in an OFDM transmission frame are known by the receiver along with the modulation techniques selected for the

carrier waves of the specific symbol, at least part of the data on the effective symbol lengths, the guard interval lengths, the number of carrier waves and the modulation techniques selected for the respective carrier waves of all the data transmitting symbols other than said specific data transmitting symbol in the frame are transmitted from the transmitter to the receiver by means of said specific symbol.

3. An orthogonal frequency division multiplexing transmission system according to claim 1, characterized in that data transmitting symbols are transmitted in such an order that the data transmitting symbols having an identical effective symbol length and a guard interval length are continuously arranged on the time base and the number of switching points where two adjacent data transmitting symbols having at least mutually different effective symbol lengths or mutually different guard interval lengths are located is minimized.

4. An orthogonal frequency division multiplexing transmission system according to claim 1, characterized in that said OFDM transmission frame comprises data transmitting symbols having a relatively long effective symbol length and a relatively long guard interval length for fixed reception and those having a relatively short effective symbol length and a relatively short guard interval length for mobile reception.

5. An orthogonal frequency division multiplexing transmission system according to claim 4, characterized in that it is used for digital television broadcasting.

6. An orthogonal frequency division multiplexing transmission system according to claim 1, characterized in that, if the average transmission power required for the i-th data transmitting symbol is Pi in the OFDM frame of the above embodiment, Pi is determined as a function of Ni that defines the effective symbol length in the frame so that Pi and Ni provide a one-to-one correspondence and that, if there are L different values Ni can take, there are also provided L different values Pi can take.

7. An orthogonal frequency division multiplexing transmission system according to claim 1, characterized in that, if the value of Ni is selected from A1, A2, ..., AL and Amax is the largest value of A1, A2, ..., AL, all the numbers Al, A2, ..., AL are so selected as to be divisors of Amax that can exactly divide the latter.

8. An orthogonal frequency division multiplexing transmission system according to claim 1, characterized in that Mi can take a single value.

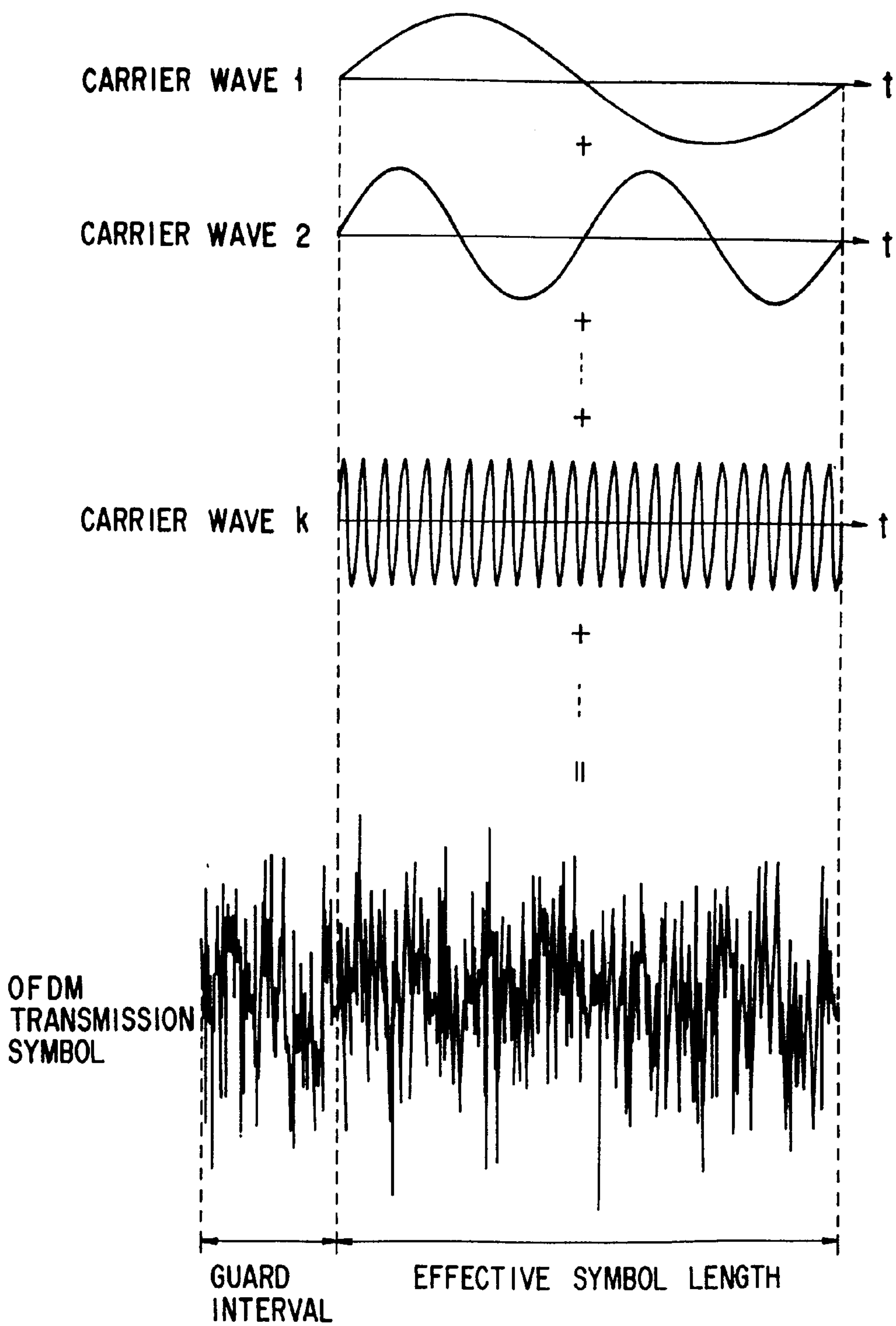
9. An orthogonal frequency division multiplexing transmission system according to claim 1, characterized in that the frequencies of the carrier waves for the data transmitting symbols in said OFDM transmission frame are shifted with a predetermined period and a predetermined frequency interval.

10. An orthogonal frequency division multiplexing transmission system according to claim 4, characterized in that at least one data transmitting symbols having a relatively short effective symbol length and a relatively short guard interval length for mobile reception are transmitted at a predetermined time interval within a frame.

11. An orthogonal frequency division multiplexing transmission system according to claim 2, characterized in that at least one of the effective symbol length, the guard interval length, the number of carrier waves and the modulation techniques is modified for each OFDM transmission frame by means of a specific data transmitting symbol on an OFDM transmission frame basis.

12. A transmitter adapted to an orthogonal frequency division multiplexing transmission system according to claim 1, characterized in that, if L FFT points are used for the modulation of data transmitting symbols in inverse discrete Fourier transform, a total of L inverse discrete Fourier transformers (121, 122, ..., 12L) are provided for each number of FFT points.

13. A receiver adapted to an orthogonal frequency division multiplexing transmission system according to claim 1, characterized in that, if L FFT points are used for the demodulation of data transmitting symbols in discrete Fourier transform, a total of L discrete Fourier transformers (421, 422, ..., 42L) are provided for each number of FFT points.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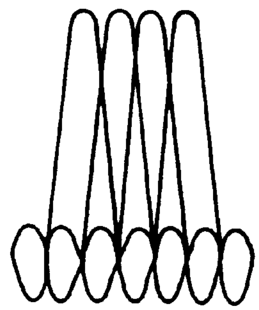
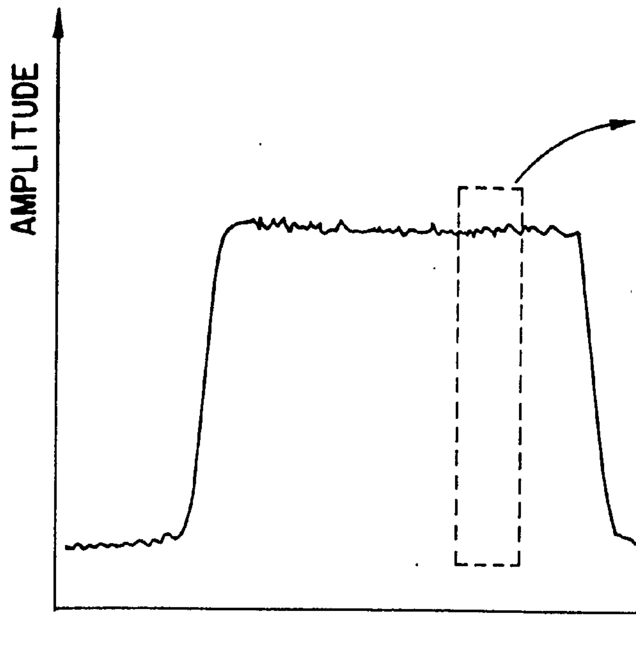
CARRIER WAVE 2
t
+
+
CARRIER WAVE k
t
+
=
OFDM TRANSMISSION SYMBOL
GUARD INTERVAL
EFFECTIVE SYMBOL LENGTH

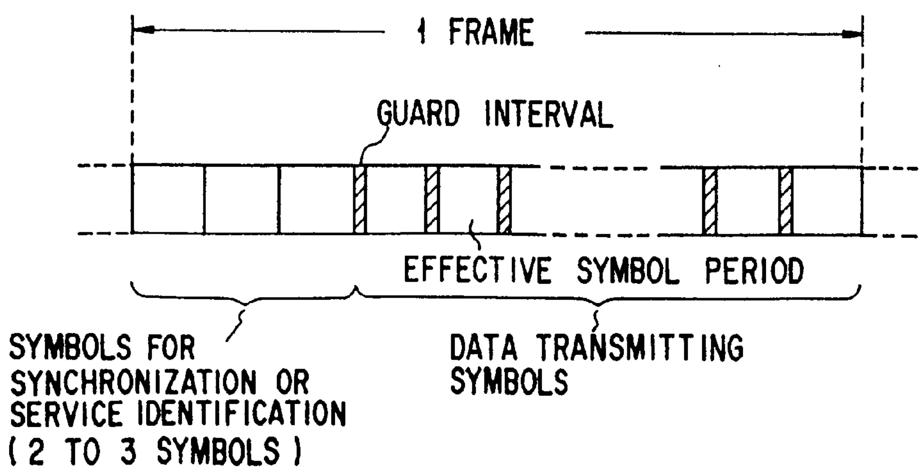
FIG. 1

F I G. 2A





F I G. 2B



F I G. 3

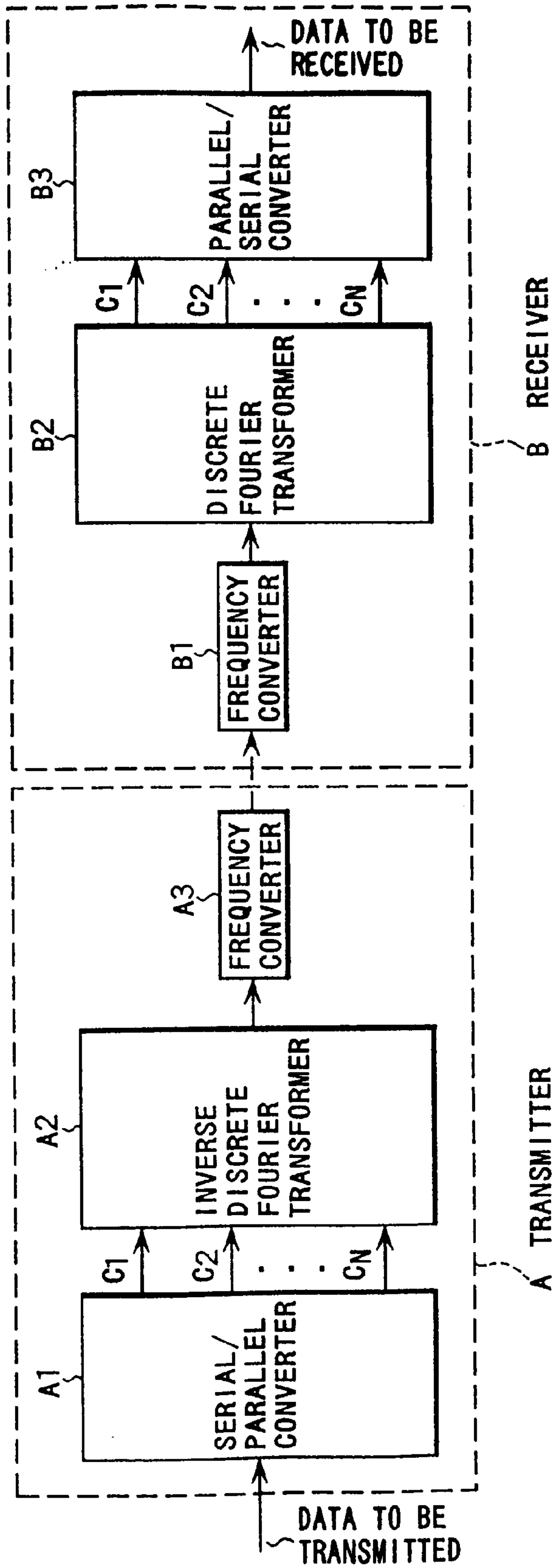
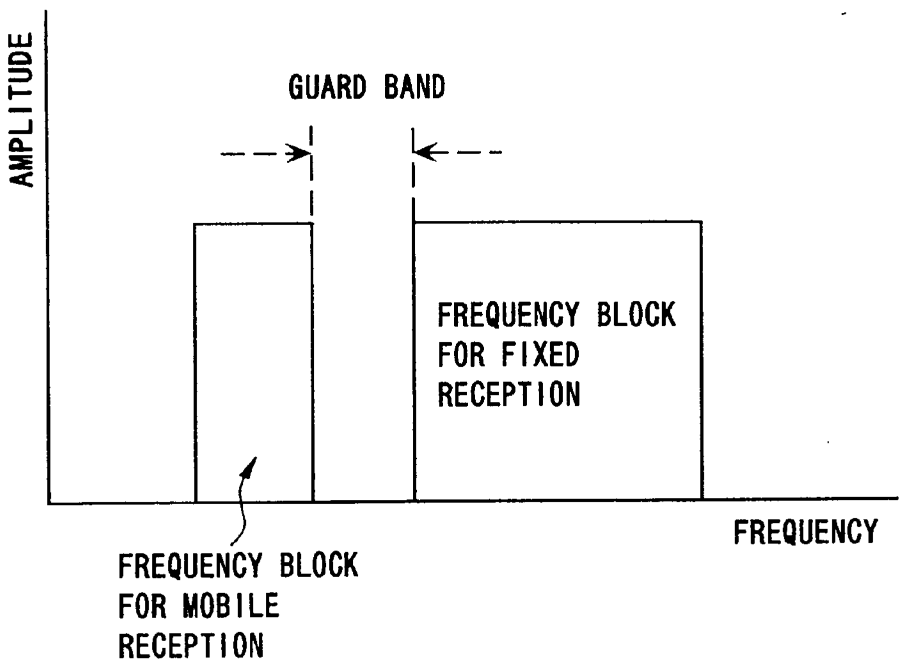
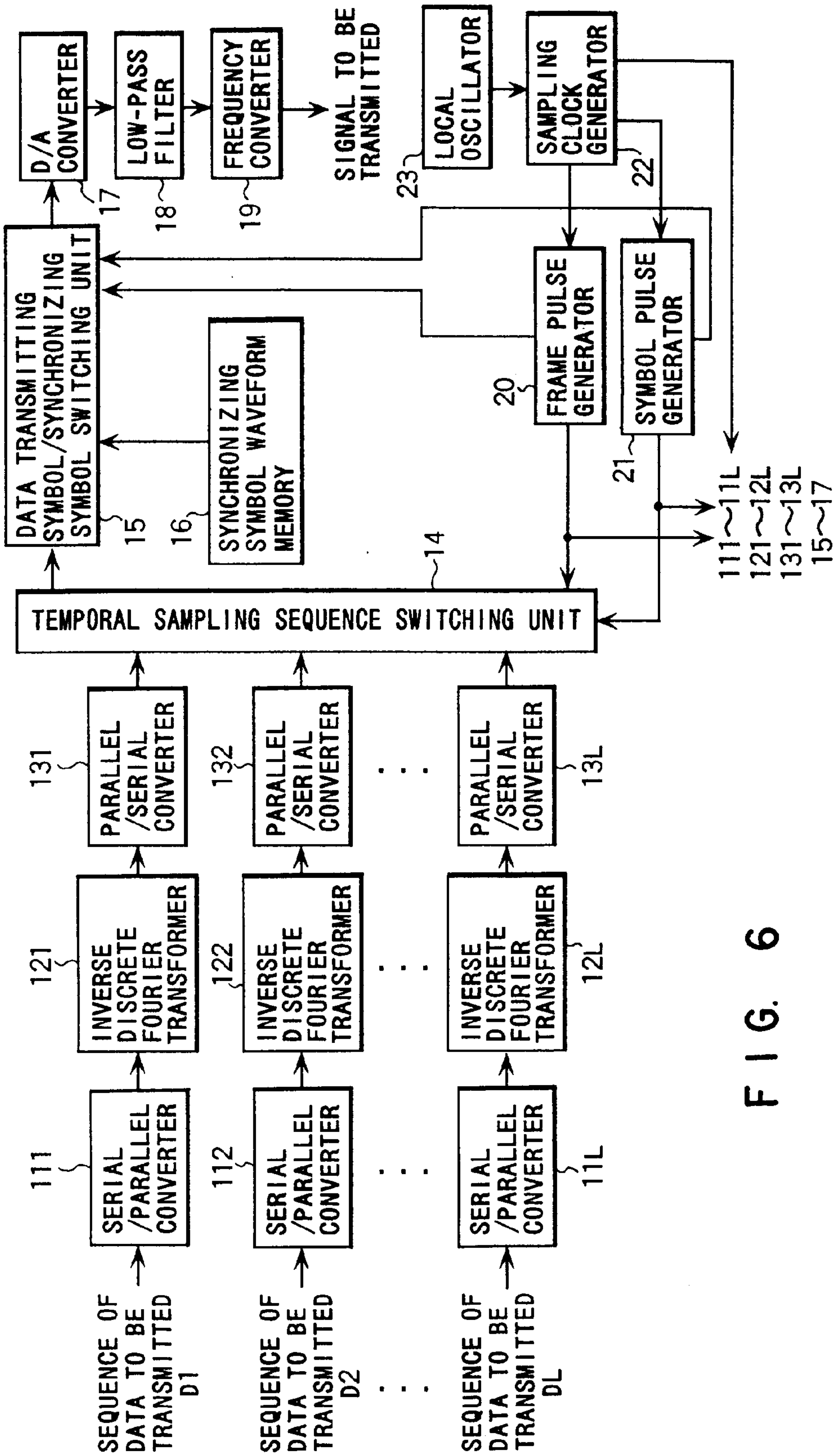

DATA TO BE TRANSMITTED
A1
SERIAL/ PARALLEL CONVERTER
C1
C2
CN
A2
INVERSE DISCRETE FOURIER TRANSFORMER
A3
FREQUENCY CONVERTER
A TRANSMITTER
B1
FREQUENCY CONVERTER
B2
DISCRETE FOURIER TRANSFORMER
C1
C2
CN
B3
PARALLEL/ SERIAL CONVERTER
DATA TO BE RECEIVED
B RECEIVER

FIG. 4


AMPLITUDE
GUARD BAND
FREQUENCY BLOCK
FOR FIXED
RECEPTION
FREQUENCY
FREQUENCY BLOCK
FOR MOBILE
RECEPTION

F I G. 5


SEQUENCE OF DATA TO BE TRANSMITTED D1
SEQUENCE OF DATA TO BE TRANSMITTED D2
SEQUENCE OF DATA TO BE TRANSMITTED DL
111
112
11L
SERIAL /PARALLEL CONVERTER
121
122
12L
INVERSE DISCRETE FOURIER TRANSFORMER
131
132
13L
PARALLEL /SERIAL CONVERTER
14
TEMPORAL SAMPLING SEQUENCE SWITCHING UNIT
15
DATA TRANSMITTING SYMBOL/SYNCHRONIZING SYMBOL SWITCHING UNIT
16
SYNCHRONIZING SYMBOL WAVEFORM MEMORY
17
D/A CONVERTER
18
LOW-PASS FILTER
19
FREQUENCY CONVERTER
SIGNAL TO BE TRANSMITTED
23
LOCAL OSCILLATOR
22
SAMPLING CLOCK GENERATOR
20
FRAME PULSE GENERATOR
21
SYMBOL PULSE GENERATOR
111~11L
121~12L
131~13L
15~17

F I G. 6

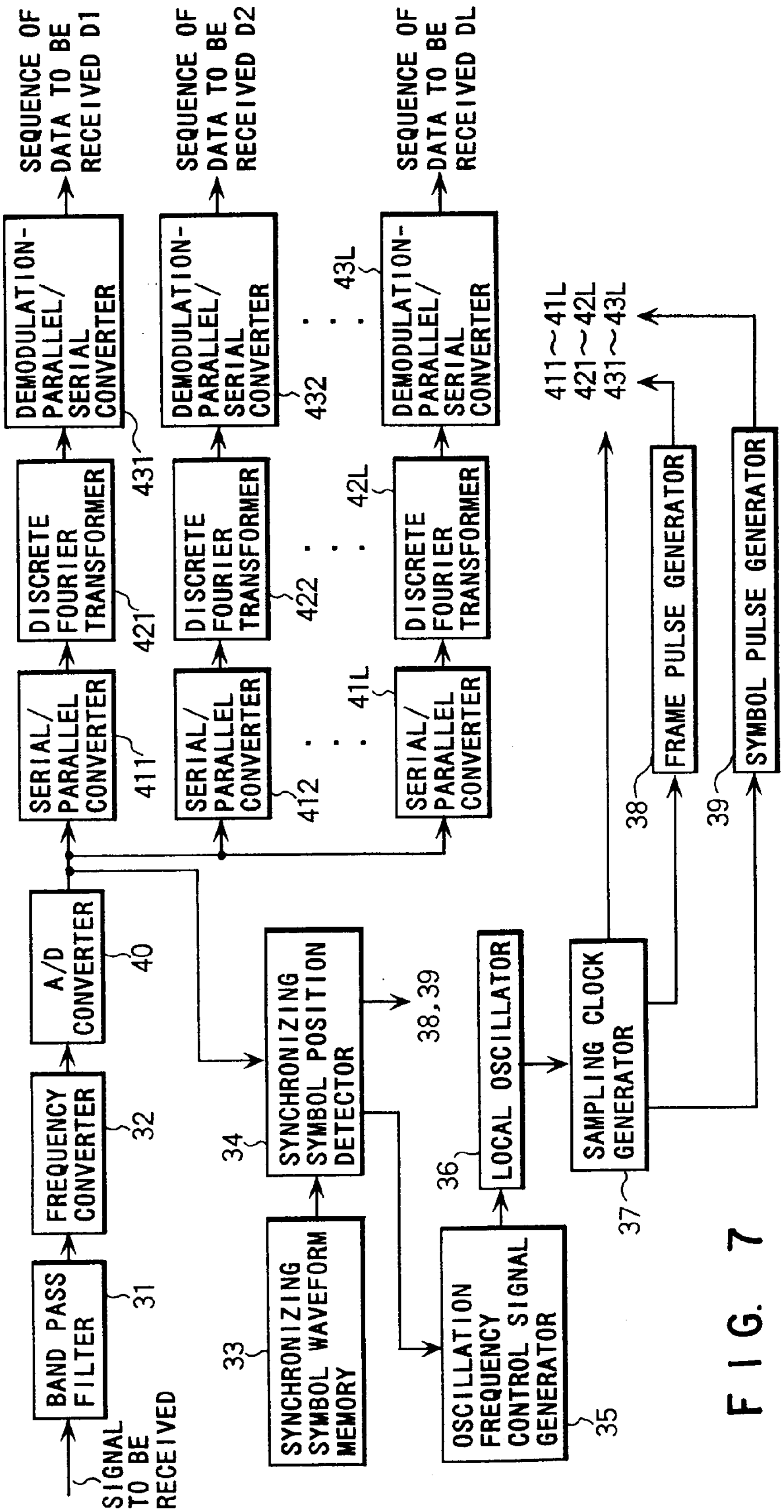

SIGNAL TO BE RECEIVED
BAND PASS FILTER
31
FREQUENCY CONVERTER
32
A/D CONVERTER
40
SERIAL/ PARALLEL CONVERTER
411
412
41L
DISCRETE FOURIER TRANSFORMER
421
422
42L
DEMODULATION- PARALLEL/ SERIAL CONVERTER
431
432
43L
SEQUENCE OF DATA TO BE RECEIVED D1
SEQUENCE OF DATA TO BE RECEIVED D2
SEQUENCE OF DATA TO BE RECEIVED DL
SYNCHRONIZING SYMBOL WAVEFORM MEMORY
33
SYNCHRONIZING SYMBOL POSITION DETECTOR
34
38, 39
OSCILLATION FREQUENCY CONTROL SIGNAL GENERATOR
35
LOCAL OSCILLATOR
36
SAMPLING CLOCK GENERATOR
37
FRAME PULSE GENERATOR
38
SYMBOL PULSE GENERATOR
39
411~41L
421~42L
431~43L

FIG. 7

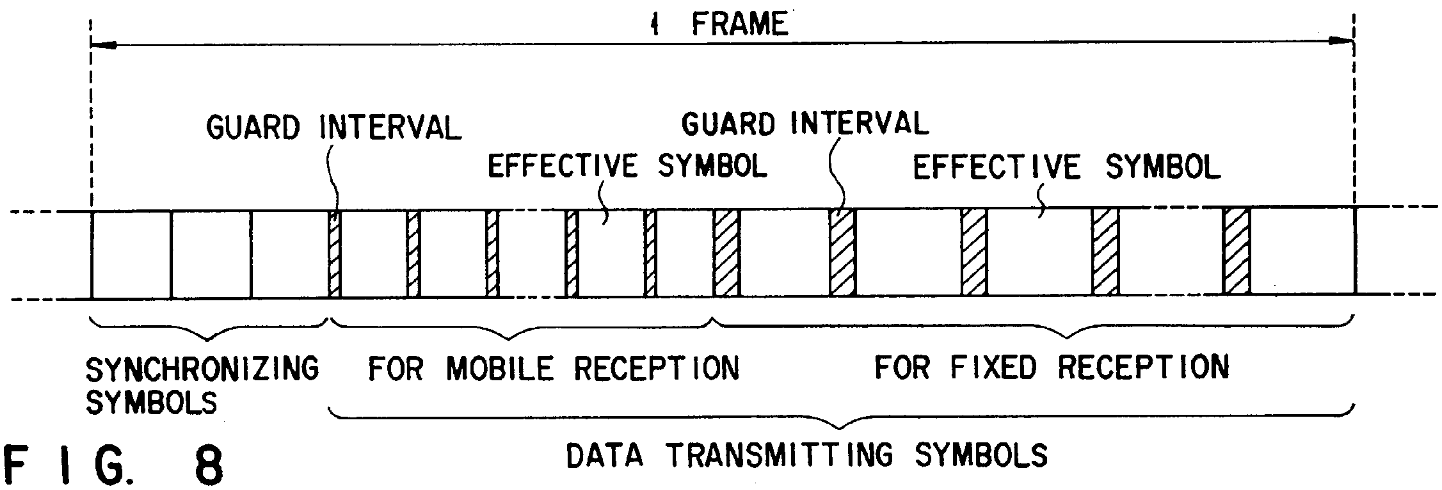


FIG. 8

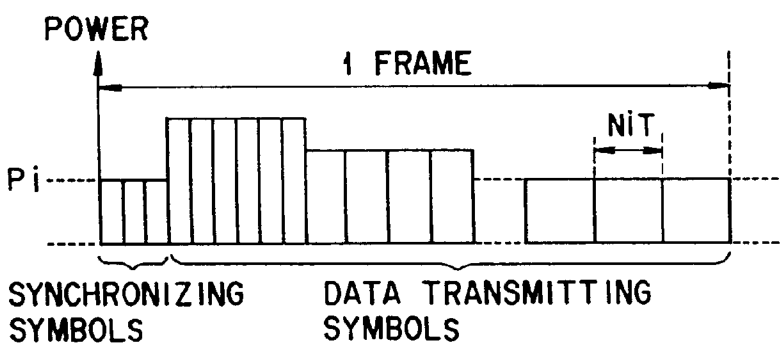


FIG. 9

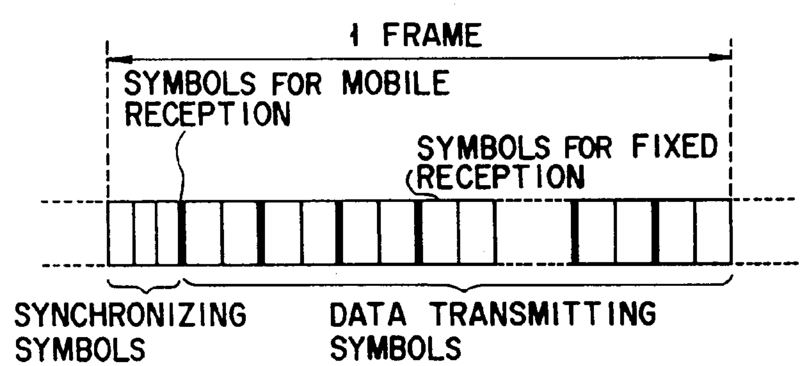


FIG. 12

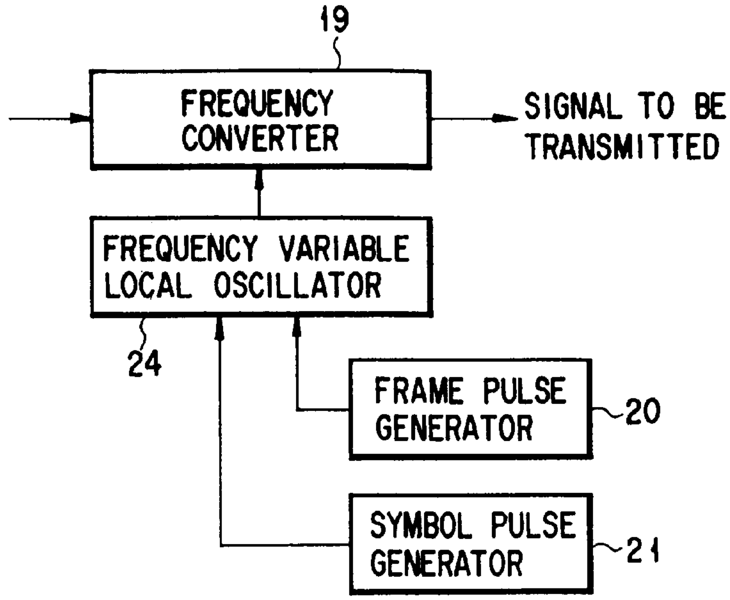


FIG. 10A

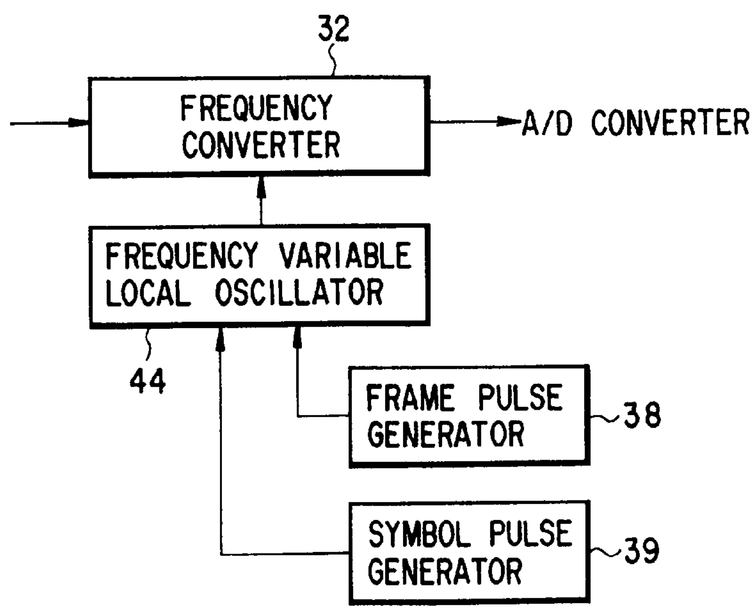
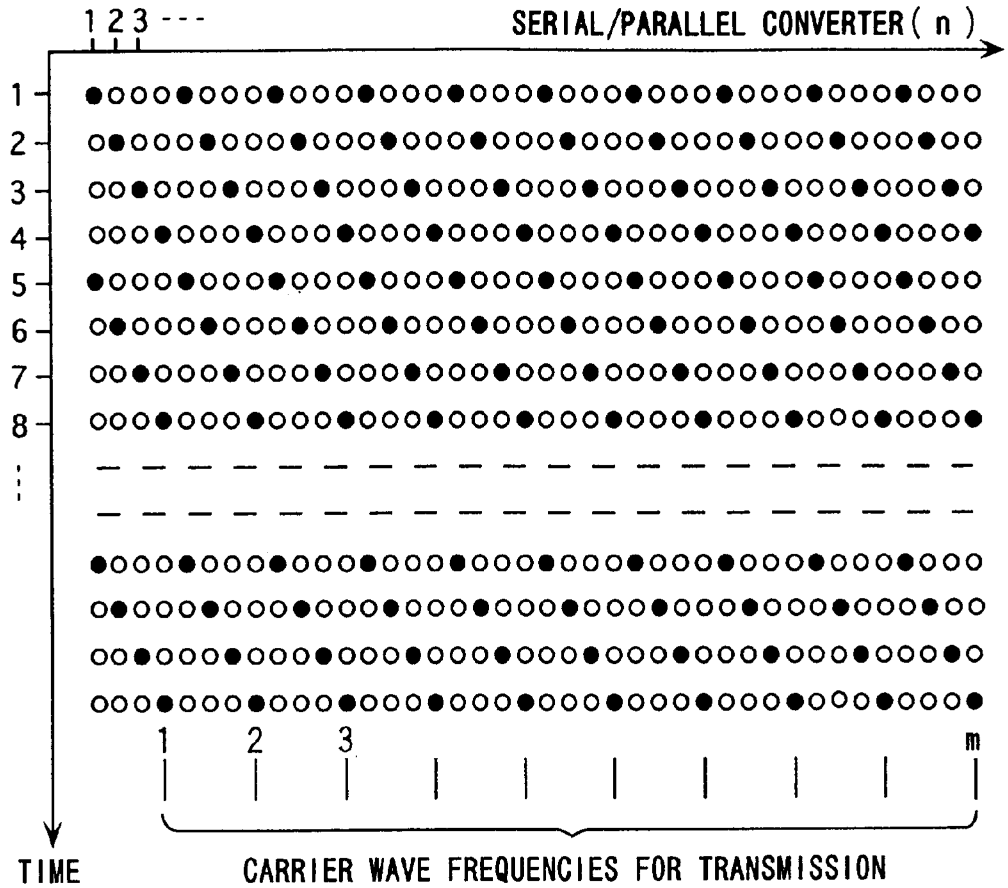
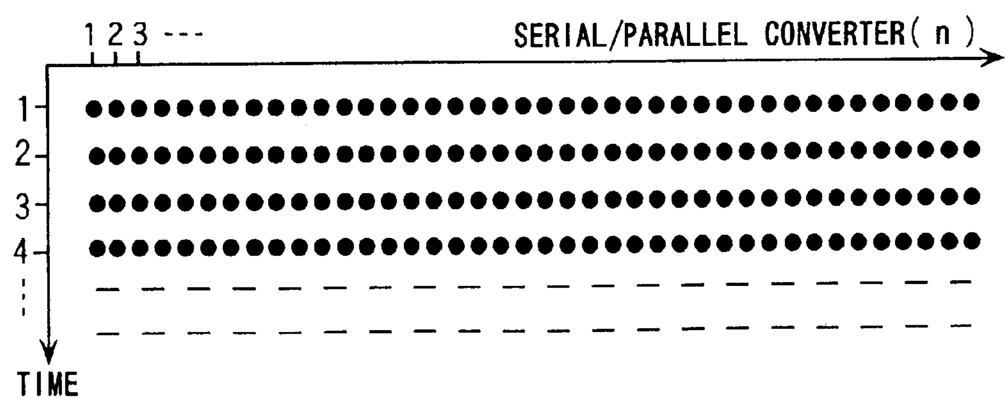


FIG. 10B



F I G. 11A



F I G. 11B

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(71) Applicant: **TRW INC.**
Redondo Beach, CA 90278 (US)

(72) Inventors:
- **Garrison, Arthur L.**
Redondo Beach, CA 90278 (US)
- **Smigla, Terrence R.**
Manhattan Beach, CA 90266 (US)
- **Stephens, Scott A.**
Manhattan Beach, CA 90266 (US)

(74) Representative:
Schmidt, Steffen J., Dipl.-Ing. et al
Wuesthoff & Wuesthoff,
Patent- und Rechtsanwälte,
Schweigerstrasse 2
81541 München (DE)

(54) Power control method and apparatus for satellite based telecommunications system

(57) A power control method and apparatus are provided for a satellite based telecommunications system. The system includes a power control subsystem which is operative with systems operations center for distributing available satellite power between earth stations. Each earth station includes a baseband manager which subdivides the available satellite power between subband beams emitted from the satellite. The earth station further includes beam processors which manage the power allocated to each subband within an associated beam in order to maintain a desired signal quality in a forward link between the satellite and user terminals within the associated subbands. The beam processors communicate with modems, each of which is assigned to a particular user terminal. Each modem controls the satellites transmission power in the forward link to the user terminals to maintain a desired signal-to-noise ratio at the user terminal receiver. The signal-to-noise ratio is determined by the corresponding beam processor. The subsystem further provides a dynamic power control loop between user terminals in the forward and return links to maintain a desired signal quality. The subsystem automatically controls the satellite output power level to ensure proper power emission by a satellite in connection with feeder links from multiple earth stations.

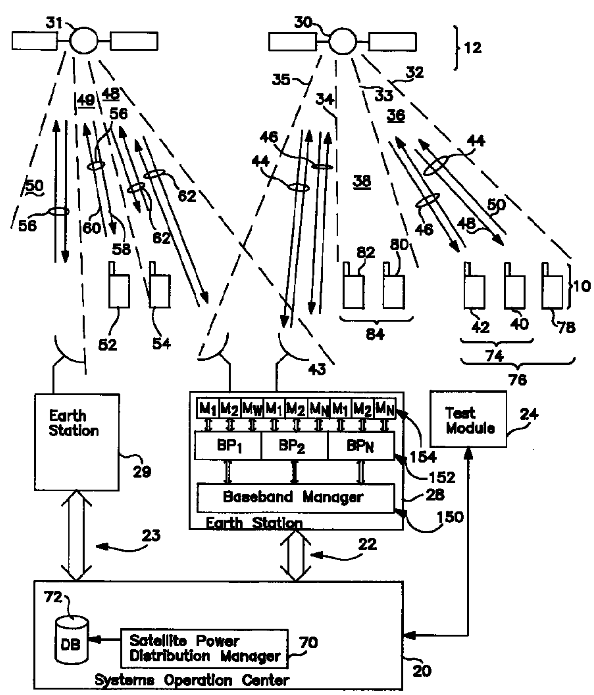


FIG. 1

EP 0 805 568 A1

Description

FIELD OF THE INVENTION

The present invention generally relates to satellite based telecommunications. More specifically, the invention relates to a power control subsystem for optimizing satellite power usage while maintaining a desired quality of service with user terminals.

BACKGROUND OF THE INVENTION

Satellite based telecommunications systems have been proposed to provide cellular communications links between user terminals (mobile and fixed) and earth stations. The earth stations, in turn, connect the user terminals with remote originating/destination callers through public land mobile networks (PLMN), public switching telephone networks, other earth stations and satellites, and the like. Each user terminal communicates with an assigned earth station along corresponding forward and return links which are supported by a satellite which has the user terminal and earth station in its field of view.

Each satellite includes at least one antenna which defines its earth coverage region or footprint. The satellite antenna(s) divide the coverage region into multiple beam spots. Each beam spot is assigned at least one frequency subband along which communications signals travel in the forward and return directions between user terminals and earth stations. Each subband may support communications from a plurality of user terminals. The user terminals are assigned unique transmission channels or "circuits" within an associated subband. A channel or "circuit" represents a unique path along which the corresponding user terminal transmits and receives RF signals containing discrete frames or packets of communications data and/or command information. A channel or circuit may be defined in a variety of ways, depending upon the system's coding technique such as time division multiple access (TDMA), frequency division multiple access (FDMA) code division multiple access (CDMA), or any combination thereof.

The transmitters in each earth station, satellite and user terminal emit an RF signal with sufficient power to ensure that the intended receiver receives the RF signal with a desired quality of service. The quality of service of a communications link is dependent on the signal-to-noise ratio (SNR) of the RF signal. Different types of user terminals (portable, fixed, special, geographically specific, etc.) have associated minimum SNR levels required to afford a desired quality of service. Thus, each satellite must transmit RF signals in associated subbands at varying power levels to maintain the desired quality of service which depend upon the intended user terminal type.

In addition, satellites vary the RF signal transmission power levels between subbands and between channels in a subband to account for system factors, such as the position of the beam spot for an associated subband, the number of user terminals assigned to the subband, the position of the user terminals within the associated beam spot, the amount of interference between the user terminal and satellite (rain, fog, clouds, etc.), the distance to the user terminal and the like. The above-noted system factors continuously change, and thus the satellite must continuously update the transmission power level of RF signals in each subband to each user terminal.

However, each satellite is afforded a limited supply of power. Each satellite has many power demands upon this limited supply. Thus, it is desirable to maximize the transmission efficiency. To do so, satellite antennas have been implemented with nonlinear amplifiers which drive the antenna array to transmit the RF signals. However, driving the nonlinear amplifiers too far into saturation will cause excessive intermodulation distortion as well as reduced amplifier reliability.

A need remains for a satellite system which optimizes the satellite transmitter operating power level, while maintaining a desired quality of service at each user terminal.

Moreover, proposed satellite systems have been unable satisfactorily control the "effective isotropic radiated power" (EIRP) emitted by an earth station and received by a corresponding satellite. As noted above, an earth station passes RF signals to a desired user terminal along a forward link of a communication channel. In the forward link, the associated satellite receives each RF signal via a feeder link with the earth station. The satellite then retransmits this received RF signal in the subband of the beam spot containing the destination user terminal. The satellite must transmit the RF signal at a power level sufficient to provide the desired signal-to-noise ratio (SNR) and quality of service at the user terminal. A need remains for a satellite system which affords control at the earth station of the power output of the satellite for each channel.

Each satellite may receive RF signals along multiple feeder links from multiple earth stations. Each earth station is located a different distance from the satellite and at a different point within the satellite field of view. Consequently, RF signals from different earth stations may be received at different power levels. Power fluctuations in the received RF signal may further vary due to signal interference, such as clouds, rain and the like. Hence, RF signals from an earth station covered by clouds would be weaker than an RF signal from an earth station with no cloud cover. A need remains for an improved feeder link between the earth stations and satellites.

The present invention provides an improved power control satellite subsystem which overcomes the disadvantages discussed above and experienced in the past.

OBJECTS OF THE INVENTION

It is an object of the present invention to provide a power control subsystem for a satellite based telecommunications system which optimally allocates power among earth stations with respect to corresponding coverage satellites.

It is a further object of the present invention to ensure that the earth stations, satellites and user terminals operate within federally mandated power flux density limits (PFD limits).

It is a further object of the present invention to provide an adjustable quality of service within forward and return communications links between earth stations and user terminals.

It is a corollary object of the present invention to enable the quality of service to be adjusted based on satellite loading, user position within the satellite's field of view, the forward link signal-to-noise ratio and terminal type.

It is yet a further object of the present invention to ensure that the power control subsystem maintains optimal control when satellite power usage approaches maximum power limits.

It is yet a further object of the present invention to initiate handover operations between beams and/or satellites to optimize satellite power load management.

It is another object of the present invention to provide an aggregate power control subsystem which distributes satellite RF signal power resources between multiple earth stations in such a way that the amplifiers driving the satellite-to-user transmitters are operated at a desired point within a nonlinear operating range to avoid signal distortion.

It is a further object of the present invention to provide a two-way user-level dynamic power control system which adjusts power transmitted to and from an individual user terminal to maintain a desired signal quality at the user terminal and at the earth station.

Another object of the present invention is to provide. automatic level control of the earth station transmission power (EIRP) and receive feeder link power at the satellite.

BRIEF DESCRIPTION OF THE DRAWINGS

Fig. 1 illustrates a block diagram of a satellite based telecommunications system according to the preferred embodiment of the present invention.

Fig. 2A illustrates a block diagram of an earth station of a preferred embodiment of the present invention.

Fig. 2B illustrates a detailed block diagram of the baseband manager and beam processor according to a preferred embodiment of the present invention.

Figs. 3A-3C illustrate the processing sequence followed by the satellite power manager of the preferred embodiment of the present invention.

Fig. 4 illustrates a more detailed block diagram of the beam processor of the preferred embodiment of the present invention.

Figs. 5A and 5B illustrate the processing sequence carried out by the beam processor of Fig. 4 according to the preferred embodiment of the present invention.

Fig. 6 illustrates a forward link power control loop between an earth station and a user terminal according to the preferred embodiment of the present invention.

Fig. 7 illustrates a return link power control loop according to a preferred embodiment of the present invention.

Fig. 8 illustrates an automatic level controller carried out according to the preferred embodiment of the present invention.

Figs. 9A and 9B illustrate exemplary RF signals transmitted in connection with the power level controller of Fig. 8 according to the preferred embodiment of the present invention.

DETAILED DESCRIPTION OF THE PREFERRED EMBODIMENTS

Fig. 1 generally illustrates a satellite based telecommunications system representative of a preferred embodiment. The system includes a plurality of user terminals 10 which communicate with corresponding earth stations 16 via coverage satellites 12. Each user terminal communicates with its assigned earth station via a unique communications channel. A channel includes a forward link from the earth station to the user terminal and a return link from the user terminal to the earth station. Each forward and return link is further divided into an earth station-to-satellite section and a user terminal-to-satellite section. Each channel carries RF signals within a preassigned subband having a central carrier frequency. Each satellite divides its coverage area (e.g., field of view) into multiple beam spots. Each beam spot may support one or more subbands. Thus, the carrier frequency of a particular channel is dependent upon the beam spot covering the user terminal. Fig. 1 illustrates an exemplary implementation of this communications architecture.

As illustrated in Fig. 1, satellite 30 divides its coverage area into three beam spots, the boundaries of which are defined by dashed lines 32-35. Beam spot 36 covers a first group of user terminals, while beam spot 38 covers a second group of user terminals. Terminals 40 and 42 communicate along channels 44 and 46, respectively, with earth station 28. Channel 44 includes a forward link 48 and a return link 50. The satellite 30 relays RF signals along channels 44 and 46 to the earth station 28.

Satellite 31 similarly includes multiple beam spots 48 and 50 which support communications between earth station 29 and user terminal 52. User terminal 52 communicates along channel 56 which includes a forward link 58 and a return link 60. As shown by channel 62, earth station 28 may also communicate with user terminals (i.e., terminal 54) which are covered by satellite 31. Earth station 28 communicates with user termi-

nal 54 via channel 62.

The earth stations 28 and 29 communicate with a system operations center (SOC) 20 through communication lines 22 and 23. The SOC 20 includes a satellite power distribution manager 70 which distributes power among the satellites 30 and 31 as explained below.

The earth station 28 includes a baseband manager 150 which communicates with SOC 20 and manages power output of the satellite 30 within subbands assigned to the earth station 28. The baseband manager 150 communicates with a plurality of beam processors 152. Each beam processor is associated with a unique beam emitted by the satellite 30. Thus, unique beam processors manage communications within corresponding beams 36 and 38. Each beam processor 152 reallocates power between channels within the corresponding beam as explained below. Each beam processor 152 communicates with a plurality of modems 154. Each modem 154 is uniquely associated with a channel and operates to maintain a desired quality of service (i.e., signal-to-noise ratio) for an associated user terminal. The outputs of the modems 154 are combined and transmitted via antenna 43 to the satellite 30 as composite RF signals.

The user terminals 10 may be of differing types, such as portable terminals, cellular terminals, fixed/stationary terminals, special terminals, geographically specific terminals and the like.

It is understood that only a subset 74 of the user terminals may be actively communicating at any given time. The system allocates a number of "radio resources" to each earth station. Radio resources represent the number of channels and/or subbands assigned to an earth station. Thus, while a group of channels or radio resources 76 may be assigned to an earth station, only the subset 74 of these terminals will be using radio resources to actively communicate.

By way of example, the system may register the user terminals in set 76 with earth station 28 since these terminals are located in a predefined fixed geographic zone of coverage assigned to the earth station. Optionally, this registration may be stored on a visitor location register stored at an earth station, at the SOC, or at a separate module which communicates with the earth stations. Alternatively, the system may calculate, through past system demands, a number of user terminals which historically have attempted to establish communications links with the earth station 28 at a given time of day. Based on these calculations, the SOC 20 informs the earth station 28 of a predicted number of channels to be needed as group 76 associated with beam spot 36. Optionally, in a CDMA or TDMA coded system, the SOC 20 may provide a number of codes usable by the earth station for each subband. The SOC further informs the earth station 28 of the number of user terminals expected to be in group 84 which will communicate along subbands corresponding to beam spot 38.

The user terminals 10 may be of differing types, such as portable terminals, cellular terminals, fixed/stationary terminals, special terminals, geographically specific terminals and the like. It is understood that only a subset 74 of the user terminals may be actively communicating at any given time. Thus, while a group 76 of user terminals 10 may be assigned to earth station 28, only a subset 74 of these terminals will be actively engaged in communications at any given time.

Satellite 30 emits RF signals along each forward transmission link in channels 44 and 46 at a power demand level determined by the earth station 28 (as explained below).

Each subband may support multiple communications channels based on any of several communication techniques, such as TDMA, FDMA, CDMA, and any combination thereof. The transmission power levels of all channels within a subband are combined to determine the power demand at the satellite for the associated subband. Each earth station is assigned to individual beam spots and operates to control the power demand for each subband within the beam spot. By way of example, a satellite 30 may cooperate with a single earth station 28 and thus, the earth station 28 controls the power demand of every beam spot for the satellite. However, when multiple earth stations 28 and 32 operate with a single satellite 31, each earth station is assigned to a subset of the beam spots for the satellite 31. For instance, earth station 28 may be assigned all user terminals in beam spot 48, while earth station 29 may be assigned all user terminals in the remaining beam spots 49 and 50. Accordingly, earth station 28 will control the power emitted by the satellite 31 within beam spot 48, while earth station 29 would control the power emitted by the satellite within the remaining beam spots.

The systems operation center (SOC) 20 is responsible for distributing the power available for use by each satellite. The systems operation center (SOC) 20 includes a satellite power distribution manager 70. The satellite power distribution manager 70 determines the total transmission power capacity of each satellite and divides the satellite's available power among the multiple earth stations communicating with the satellite. The satellite's total available power and distribution per earth station may be determined in a variety of manners, such as from preassigned values for each satellite stored in a data base 72. Optionally, the power distribution manager 70 may empirically calculate the power distribution between earth stations for a satellite based on historical use data evidencing past demands of the satellite over a desired period of time. As a further alternative, the power distribution manager 34 may periodically recalculate the total power available to each satellite based on power feedback reports from each earth station. As yet a further alternative, the power distribution may be based on the number of user terminals registered in a zone of coverage. The system's overall power may be shifted to optimize or equally load each satellite by reassigning user terminals and/or subbands between satellites and/or earth stations.

The SOC 20 informs each earth station 28 and 29 of the available satellite power which may be used in connection with user terminals assigned to the earth station. For instance, the SOC 20 may inform earth station 28 that it may distribute 500 watts between the beams/subbands of satellite 30 which are assigned to the earth station 28, and 200 watts among the beams/subbands of satellite 31 which are assigned to earth station 28. Thus, earths station 28 may instruct satellite 30 to transmit up to 500 watts of transmission power within the subbands assigned to earth station 28. Similarly, the earth station 28 may instruct satellite 31 to transit up to 200 watts of power in the beams/subbands assigned to earth station 28. In addition, the SOC 20 may inform each earth station of a maximum power limit which may be transmitted by each satellite per subband per beam spot. This power limit is determined by the SOC 20 in order to ensure that the overall system does not exceed the power flux density regulatory requirements as established by the Federal Communications Commission.

Upon receiving the satellite power allocations and regulatory limits from the SOC 20, each earth station thereafter independently controls the power levels of RF signals transmitted by satellites to each user terminal. As explained below, the baseband managers 150 in the earth stations distribute satellite transmission power among the predicted channels without exceeding the subband regulatory power limits and the satellite's available power assigned to the corresponding earth station.

Throughout operation, each earth station provides power demand feedback information to the SOC 20 which is used to update the power allocation among the earth stations. By way of example, the feedback information may include the total power required of the satellite to maintain communications links with minimum signal quality. In this manner, the SOC 20 monitors the actual and required satellite power usage relative to ideal operating power levels. The SOC 20 periodically updates the satellite power allocations to each earth station based on feedback information from the earth stations concerning loads and required satellite transmitter operating power levels.

Optionally, a mobile link test module 24 may be provided for measuring a satellite transmitter operating level. The test module 24 communicates measurements directed to the SOC 20. Alternatively, or in addition, a telemetry channel may be maintained between the satellite and each associated earth station. When the telemetry channel is used, the satellite may telemeter transmission operating information to the earth station which in turn relays it to the SOC 20. The SOC 20 in turn utilizes the telemetered satellite operating information while updating the power allocations.

Each earth station estimates its current total satellite power usage relative to the allocated power. Each earth station estimates and controls its satellite power usage per subband per beam relative to the regulatory power limits provided by the SOC 20. Each earth station performs user level power control and dynamic fade margin adjustments for each user (as explained below). Periodically, the earth stations report total satellite power usage, along with power usage per subband per beam for each associated satellite.

Turning to Fig. 2A, an earth station 28 is illustrated in more detail. The earth station includes a baseband manager 150, a plurality of beam processors 152, and a plurality of modems 154. Each beam processor operates in connection with an assigned beam emitted by the satellite. Each beam processor includes one or more subband power managers 156 which manage power distribution among the subbands in the associated beam. Each subband power manager 156 communicates with a plurality of modems 154. Each modem 154 operates in connection with a single channel assigned to a particular user terminal. Each subband power manager 156 communicates with all of the modems 154 which support channels in a single corresponding subband. Each modem 154 includes a forward link power controller 160 which controls the power emitted by the satellite within the subband corresponding to the channel assigned to the modem 154. Each modem 154 includes a modulator 162 and a demodulator 164 for modulating and demodulating RF signals transmitted from and received by the earth station in connection with the associated channel. The RF signals emitted by modulators 162 within modems 154 corresponding to a single subband are combined at a summer 166 prior to transmission to form a composite RF signal for the subband. The composite RF signals are transmitted along with a reference tone (explained below).

The baseband manager 150 includes a satellite power manager 158 and beam load manager 161 which operates according to the flow process illustrated in Figs. 3A-3C to control power distribution among beams transmitted by the satellite to user terminals assigned to the earth station.

Fig. 2B illustrates the interconnection between the baseband manager 150 and a beam processor 152 in more detail. The satellite power manager 158 receives the total satellite power allocation for the earth station from the SOC. The satellite power manager 158 also receives the number of expected channels to be assigned to the earth station per beam of the associated satellite. The satellite power manager 158 receives, as feedback, the difference between the required and allocated total power per subband for each beam from the beam processor 152. The satellite power manager 158 communicates with the data base 153 which may store PFD limits downloadable from the SOC. The PFD limits may be accessed by geographic region and carrier frequency which are dependent upon the satellite's current position and the beam of interest. Referring to Fig. 3A, the satellite power manager 158 obtains (step 170) the total available satellite power from the SOC 20. At step 172, the satellite power manager 158 obtains the subband PFD limits and at step

174 obtains the channel assignments per beam (from the SOC 20 or a database). At step 176, the satellite power manager determines the power to be allocated to each subband.

This determination may be based on the number of potential user terminals expected to request channels within a particular subband. Alternatively, this determination may be based on the position within the satellites field of view of the beam containing the present subband. As a further alternative, the subband power allocations may be based on data concerning past usage demands. By way of example, user terminals may register according to one of several processes set forth in co-pending applications entitled "Satellite Based Cellular Telecommunications System Utilizing A Multiple Registration Location Register" and entitled "Earth Stationed Subsystem" filed on or about May 1, 1996 and assigned to the assignee of the present invention. The two above noted co-pending application are incorporated herein by reference in their entirety. At step 178, the subband power allocations are passed to corresponding subband power managers 158.

At step 180, the satellite power manager 158 obtains the feedback differential between the subband required power and the subband allocated power from the Subband Power Manager 156. At step 182, the satellite power manager 158 combines the feedback differentials for all of the subbands of the current satellite and reports, to the SOC, the total required power and the total allocated satellite power. Next, the satellite power manager 158 reports to the beam load manager 161 (Fig. 2B) the required subband power and the allocated subband power. Next it determines that step 186 (Fig. 3B) whether the allocated subband power exceeds the required subband power. If so, control passes to step 190 at which the satellite power manager 158 records the amount of excess subband power for future use by other subbands which may need additional power. Returning to step 186, if the allocated subband power does not exceed the required power, then the subband needs additional power. Hence, flow passes to step 188 at which the satellite power manager 158 determines whether other subbands have recorded excess power which may be reallocated to the present subband in need of additional power. At step 192, if such additional power exists, the satellite power manager 158 reallocates the power from the subband in excess to the subband in need and passes new power allocations per subband to the subband power manager 156. Excess power represents the power allotted to a subband, but not needed to achieve the desired quality of service for the user terminals currently in use. The excess power report also indicates when the user terminals demand more power from a particular beam than has been allocated thereto by the satellite power manager 158. The satellite power manager 158 uses the excess power report to reallocate power between beams, such as when one beam requires more power than has been allocated thereto, while an adjacent beam does not require all of its allotted power.

When the excess power reports for all of the beams assigned to the earth station indicate that additional power remains, optionally, the satellite power manager 158 may distribute the excess additional power among the subbands. The distribution of additional power may be even or uneven according to some other desired function. Once the manager 158 allocates all of the available power to the beam spots, the baseband manager 150 reports back to the SOC 20 the difference between the necessary minimum total satellite power and the allotted power. Thus, if the beams assigned to the earth station 28 only require 80% of the total power allocated by the SOC 20, the baseband manager 150 returns this information to the SOC 20. Optionally, the SOC 20 may shift the unneeded 20% to beams upon the satellite which have been assigned to a different earth station by outputting new satellite power allocation amounts.

According to the foregoing process, the baseband manager 150 continuously updates the subband power allocations for each beam based on the total satellite power allocation, the PFD limits and the usage recorded from each beam processor 152.

Turning to Fig. 4, a beam processor 152 is illustrated in more detail. The beam processor 152 includes a subband power manager 156 which includes a signal-to-noise (SNR) ratio calculation module 157. The SNR calculation module 157 accesses a data base 155 (Fig. 2) to obtain a required SNR value for the current user terminal based on the user terminal's type. The user terminal's type may be provided by the baseband manager 150, or may be stored in data base 155. The SNR value obtained from data base 155 represents a minimum required SNR value necessary to achieve a quality of service desired by the user terminal. The SNR calculation module 157 further receives the forward link SNR variance from the modem assigned to the current user terminal. The SNR calculation module 157 also obtains a "fade margin" for the current channel associated with the current user terminal. The "fade margin" represents a predetermined bias value added to the minimum required SNR value for a user terminal in order to compensate for rapid fluctuations in the received power level at the user terminal. These fluctuations are sufficiently rapid that they are difficult to correct by the system through its normal power control loop. Thus, a bias or "fade margin" is added to the minimum SNR value to ensure that, during a minimum of a rapid fluctuation, the power level never falls below the lowest acceptable value. The beam processor may determine the fade margin for each user terminal based on the subband power allocation received from the satellite power manager 158, the user terminal type and position within an associated beam spot. The fade margin may also be determined based on forward SNR feedback statistics reported from a modem 154 corresponding to the current user terminal. The fade margin may also be based on the current power usage of the modem. The SNR

calculation module 157 combines the required SNR value, SNR variance and fade margin to generate a new required SNR value for the forward link (FL) with the current user terminal. This required FL SNR value is supplied to the modem control module 159.

Figs. 5A and 5B illustrate the processing sequence carried out by the subband power manager 156. Initially, modem controller 159 obtains the subband power allocation from the satellite power manager 158 for the current subband (step 200). At step 202, the SNR calculation module 157 calculates the required SNR value as described above. The modem controller 159 outputs the desired SNR value to the modem 154 corresponding to the current channel within the current subband. The modem controller 159 also outputs the power limit which may not be exceeded by the modem. The modem 154 drives the satellite to emit sufficient power in the forward link to establish the desired SNR value at the user terminal. The modem thereafter returns the power level required of the satellite to achieve the desired SNR value. The modem controller 159 receives feedback information from all of the modems corresponding to the current subband and determines the total required power of the current subband to achieve the desired SNR values for each channel within the current subband. The modem controller 159 then determines whether the allocated power for the current subband exceeds or is less than the total power required to achieve the desired SNR values for each active channel within the subband. The modem controller 159 distributes this excess power by determining a desired forward link (FL) SNR value for each user terminal. The desired FL-SNR value represents the SNR level to be maintained by each modem for the forward link of the associated channel. The modem controller 159 calculates the desired SNR level for the current modem based on the associated user terminal's desired FL-SNR value and the excess power available (step 206 in Fig. 5A). The modem controller 159 outputs the desired FL-SNR value and outputs the maximum power level to which the modem may drive the satellite transmitter for the associated channel.

As explained below, each modem 154 continuously adjusts the output power of its associated channel to maintain the desired received SNR value in the presence of beam spot motion and user motion. Thereafter, the modem returns, to the modem controller 159, the forward link satellite output power level emitted in the current channel by the satellite. At step 208, the modem controller 159 combines FL required modem power levels returned from each modem for a subband to determine the total subband power. The modem controller 159 obtains a difference between the required FL subband power and the available FL subband power allotted by the baseband manager and returns a difference power level to the satellite power manager 158 (step 210). The subband power difference represents the difference between the available subband power, as provided by the baseband manager, and the required subband power, as determined by the feedback, from the modems for the current subband.

Turning to Fig. 5B, once the subband power difference is calculated at step 210, flow passes to step 212 at which the controller 157 determines whether the allocated subband power exceeds the required subband power. If so, the excess power is distributed among the modems in a desired manner (step 214). In addition, this excess is reported back to the satellite power manager 158. As explained above, the satellite power manager 158 may decide to take away the excess power from the current subband and allocate it to another subband and/or beam (see Figs. 3A and 3B). If the decision in step 212 is negative, the flow passes to step 216 at which it is determined whether the required subband power exceeds the allocated subband power. If so, the modem controller 159 reduces the desired FL-SNR values for the user terminals within the current subband in order that the output power level associated with the desired FL-SNR values does not exceed the allocated power level.

Optionally, the desired FL-SNR value for each user terminal may be reduced unevenly across the subband such as to maintain the desired FL-SNR value of each user terminal by a proportional amount above the minimum required FL-SNR value for each user terminal. This overpowered condition is reported at step 218 to the satellite power manager 158 which will subsequently, if possible, allocate additional power to the subband in an overpowered state (as explained above in connection with Figs. 3A and 3B). In addition, at step 220, the subband load manager 163 may be instructed to direct new calls to and from user terminals in the same beam to another subband other than the current subband which is operating in an overloaded power state. Thus, the subband load manager 163 distributes new calls among the subbands in order to avoid overloading of a single subband. The subband load manager 163 may operate independently in response to the feedback reported from the subband power manager 156 or alternatively under the direct control of the beam load manager 161 in the baseband manager 150.

At step 192 (Fig. 3B), the beam load manager 161 may determine which subbands within the current beam use the least power to direct the subband load manager 163 to redirect new calls accordingly. The beam load manager 161 then assigns a new channel to this underpowered subband and relays the channel assignment to the subband load manager 163. The subband load manager 163 then uses this assignment information to establish a new channel with the new user terminal.

Optionally, at step 222, a handover processor 165 within the subband load manager 163 may be activated to handover one or more active channels from the current subband to a different subband within the same beam spot. By handing over channels between subbands in this matter, the handover processor 165 shifts load between subbands. The handover processor 165 may be controlled by the subband power manager 156

and/or by the beam load manager 161. If controlled by the subband power manager 156, the handover processor 165 receives its instructions to effect a handover at step 222 (Fig. 5B). If the handover processor 165 is controlled by the beam load manager 161, the handover instructions will be transmitted at step 192 (Fig. 3B) as part of the subband power reallocation process carried out by the baseband manager 150.

According to the above process, each beam processor 152 receives, from the satellite power manager 158, the total power allocated for each subband within the associated beam. The beam processor 152 receives, from each modem 154, the current power output level for each forward link to an active user terminal, along with the variance within the forward link's signal-to-noise ratio. The signal-to-noise ratio variance represents the statistical difference in the SNR value received at the user terminal and the desired SNR value assigned to the modem 154. The beam processor 152 may receive from each modem 154 the type of user terminal being serviced by the modem 154 in an active communications link.

The beam processor 152 outputs a desired FL-SNR value to be received at the user terminal, along with a maximum satellite power output in connection with each modem 154. The beam processor 152 returns to the satellite power manager 158 a difference between the available subband power and the required subband power for each subband within the beam associated with the current beam processor. The beam processor 152 may access a data base 155 to obtain required FL-SNR values for each type of user terminal and any other desired SNR statistics, such as high variance, low variance, etc. The data base of SNR values is downloadable from the baseband manager each time the beam processor is assigned to a new beam. Thus, the content of the data base 155 may be a function of the current beam's position within the satellite's field of view.

Fig. 3C illustrates the processing sequence followed by the satellite power manager 158 when it determines that the total demand of all subbands in all beams assigned to the present baseband manager (earth station) exceed the satellite power allocation for these beams (step 194). At step 196, the satellite power manager 158 reduces the subband power allocations for all beams associated with the current earth station. This reduction may be performed according to a predefined fixed threshold value for each subband. Alternatively, this power reduction may be performed according to a predefined or calculated percentage of the total power allocated to each subband in order to reduce the subbands more evenly. At step 198, the satellite power manager 158 informs the SOC that the power demand has exceeded the allocated power for the beams corresponding to the current earth station. As explained above, the SOC may redistribute power allocation between earth stations to more evenly load the satellites and beams therein. Optionally, the SOC may also reassign user terminals and/or subbands and/or beams between overlapping satellites to shift load from the satellite operating in an overloaded position to a satellite operating in an underloaded state.

Next, the discussion turns to the power control loops used in connection with the forward link FL (Fig. 6) and the return link RL (Fig. 7) of a channel to ensure a desired signal quality at the user terminal 10 and at the earth station 20, respectively.

Beginning with Fig. 6, an earth station 28 and a user terminal 10 are illustrated. While the intermediate satellite has not been illustrated, it is understood that the communications links between the earth station and user terminal 28 and 10 pass through an associated coverage satellite. The earth station 28 includes a transmitter 11 which transmits RF signals along a forward link FL to a receiver 2 at the user terminal 10. The receiver 2 passes the incoming RF signal to a signal processor 4 which determines the received signal quality (e.g., SNR). The signal processor 4 outputs a signal-to-noise ratio (SNR) value corresponding to the received RF signal. The SNR value is combined in a multiplexor 6 with an outgoing traffic signal which is passed to the transmitter 8 and transmitted to the earth station 28 via an RF return link RL. A receiver 9 at the earth station 28 receives the RF signal upon the return link RL. The RF signal is passed to a modem 154 which demodulates the RF signals and demultiplexes the SNR value from the traffic information. The modem 154 compares the received SNR value with the desired SNR value (delivered from the corresponding beam processor 152). Based on this comparison, the modem 154 increases or decreases a power level supplied to the transmitter 11. As explained below the power level setting instructs the satellite to increase or decrease the output transmission power of subsequent RF signals along the associated forward link.

The power level control loop of Fig. 6 is repeated continuously throughout communication between an earth station and each active user terminal in order to maintain the output power of the satellite at a level sufficient to ensure that the received SNR value at the user terminal 10 substantially corresponds to the desired SNR value determined in the earth station 28. Optionally, the signal processor 4 may be modified to calculate the received SNR value based on several incoming RF signals in order to obtain an average of these multiple received SNR values. By averaging the received SNR values for multiple incoming samples, the processor 4 avoids unnecessary drastic short term variations in the output power level.

With reference to Fig. 7, the power level control loop is now described in connection with the return link RL to ensure that the receiver 9 in the earth station 28 receives the desired signal quality. Beginning at user terminal 10, an RF signal is output by transmitter 8 along the return link RL which is received at receiver 9. An SNR test module 13 tests the signal-to-noise ratio of the incoming RF signal at receiver 9. The received SNR

value is compared with the desired SNR value, and the difference therebetween is used to determine a new power setting command to be passed to the user terminal 10. The new power setting command identifies the power level at which the transmitter 8 must emit RF signals along return link RL to ensure that the satellite receives such RF signals with sufficient quality. The power setting commands are combined with an outgoing traffic signal within a multiplexor 7 and passed to the transmitter 11. The transmitter outputs the RF signal containing the power setting commands along the forward link FL to the terminal 10. A demultiplexor 3 separates the power level commands from the traffic signal and passes the power level commands to the transmitter 8. The transmitter 8 updates its output power based on the received level command. According to the foregoing loop, the return link power is maintained at a desired level.

With reference to Fig. 8, next the discussion turns to the process used to automatically control the power output levels of the satellite transmitter in forward links to all of the associated user terminals. Fig. 8 illustrates a satellite 300 which receives RF signals transmitted by earth stations 302-306 along forward feeder links 308-312. Each earth station 302-306 includes a baseband subsystem 314 which communicates with an antenna subsystem 316. The baseband subsystem 314 includes a multiplexor 318 which receives RF signals containing communications data, command information and the like along traffic channels 320 for all of the user terminals assigned to the earth station 302. The multiplexor 318 combines the RF signals along traffic channels 320 with a reference tone produced by tone generator 322. The communications signals and reference tone are passed along line 324 to the antenna subsystem through an automatic gain controller 326.

The automatic gain controller 326 is controlled to adjust the aggregate output power transmitted by antenna 328 along the feeder link 308. The RF signal transmitted along forward link 308 is received at a feeder link 330 and passed to an automatic gain controller 332. The gain of the automatic gain controller (AGC) 332 is adjusted to force the level of the reference tone embedded in the RF signal to achieve a desired level of the AGC output. By adjusting the gain at AGC 332, the reference tones from each of the multiple feeder links are driven to the same power levels while maintains the relationships between individual user power and the reference tone. In this way, any differences in propagation loss between the multiple feeder links have been compensated prior to combining the RF signals. The 3-way combiner 336 combines the RF signals received at feeder links 330, 338 and 340, respectively, and outputs same from the antenna 342 which defines the coverage region of the satellite. Next, an example is illustrated in connection with Figs. 9A and 9B to explain the manner in which the preferred embodiment achieves automatic level control.

Fig. 9A illustrates an exemplary RF signal 350 produced by the multiplexor 318. The RF signal 350 includes communications data for multiple subbands 352, 354 and 356. The composite signal 350 also includes a tone 358 produced by tone generator 322. The reference tone 358 has an amplitude corresponding to a predefined power output level. For instance, the tone 358 may correspond to two watts of transmission power ultimately transmitted by the satellite 300. The composite RF signal 350 is passed through the antenna subsystem 316 and transmitted from antenna 328.

During transmission, the RF signal may pass through interference, such as clouds, rain and the like. Such interference may alter the magnitudes of the signals within each subband 352-356 and the magnitude of the reference tone 358. The received composite signal 360 in Fig. 9A is representative of the signal received at feeder link 330. The received composite signal 360 includes subband signals 352-356 and a reference tone 358. The magnitudes of the subband signals and reference tone have increased, although, the relative amplitudes between the subband signals 352-356 and the reference tone 358 have not changed. The filter 334 adjusts the gain of the AGC 332 until it outputs the received reference tone 368 at an amplitude corresponding to the predefined amplitude associated with the predetermined output power level (e.g., two watts). Thereafter, the AGC 332 is controlled to output the adjusted composite RF signal 370 (Fig. 9A). As adjusted by the AGC 332, the RF signal 370 includes a referenced tone 378 equal in magnitude to the original reference tone 358 output by the multiplexor 318. In addition, the amplitudes of the RF signals in subbands 372-376 equal the amplitudes of the original subband signals 352-356.

Accordingly, by combining reference tones preassigned to a corresponding transmission power level, the earth station is able to ensure that the satellite receives over the feeder link a composite RF signal having a desired relation between the tone and traffic signals transmitted from the earth station. The subsequent signal transmitted from antenna 342 corresponds in amplitude to the amplitudes established by the relation between subband signals 372-376 and reference tone 378. Accordingly, by adjusting the amplitude of the subband signals 352-356 at the modems relative to the reference tone 358, the earth station is able to control the transmission power generated within each subband at the antenna 342.

Fig. 9B further illustrates a second example of the automatic level control process according to the preferred embodiment of the present invention. Fig. 9B illustrates an original composite RF signal 380, a received composite RF signal 390, and an adjusted composite RF signal 400. The original and adjusted composite signals 380 and 400 include subband signals 382-384 and 402-404, which are equal in amplitude. Reference tones 386 and 406 are also equal in amplitude. This amplitude relationship is maintained even though the received composite RF signal 390 included

subband signals and a reference tone 392, 394 and 396 which substantially differed in amplitude from the original intended signal.

The adjusted composite RF signals 270 and 300 are combined in the 3-way combiner 236 and transmitted along corresponding subbands to the intended terminals.

According to the above described preferred embodiment of the present invention, the SOC 20 allocates total satellite power and delivers same to each earth station. Within the earth station, the baseband manager subdivides the satellites total allocated power between beams and subbands. The beam processors, upon receiving the subband power allocations, control sets of modems associated with each subband to achieve required forward link SNR values. The modems adjust the output power of RF signals transmitted from the antenna subsystem to the satellite via the feeder link. The modems maintain a desired relation between subband power levels and a reference tone within each transmitted RF signal. The satellite adjusts the AGC associated with each feeder link based on the reference tone within each received RF signal to properly adjust the gain of the incoming RF signal. The satellite then combines the adjusted RF signals from the multiple feeder links in a combiner and transmits a composite RF signal from the antenna throughout the satellites field of coverage with corresponding subbands of predefined power within each beam spot. The modems receive feedback with respect to the signal quality and adjust the satellites output power by adjusting the amplitudes of the RF signals within the composite RF signal 350. The beam processors, baseband manager and SOC adjust the power allocated to each beam based on feedback from the modems.

Optionally, a test module 24 may be provided proximate the earth station to receive a test RF signal transmitted from the satellite 30. The test module 24 measures the total feeder link power relative to the reference tone based on a return signal from the satellite 30.

Optionally, the baseband manager and beam processor may cooperate to redistribute power initially reserved for future predicted channels to active channels which require additional power to achieve the minimum acceptable SNR value. For instance, if a particular subband is predicted to include 20 channels, the baseband manager and beam processor will reserve sufficient power to service 20 channels if they become active. However, during processing if 5 channels are active and the remaining are inactive, the beam processor and baseband manager may reallocate a percentage of the reserved power to other beams and/or subbands which are in need of additional power. The baseband manager and beam processor will initially divide the available power to support a predetermined minimum number of additional active channels from the 20 predicted channels.

While particular elements, embodiments and applications of the present invention have been shown and described, it will be understood, of course, that the invention is not limited thereto since modifications may be made by those skilled in the art, particularly in light of the foregoing teachings. It is, therefore, contemplated by the appended claims to cover such modifications as incorporate those features which come within the spirit and scope of the invention.

Claims

1. A power control subsystem in a satellite based telecommunications system for distributing satellite transmission power among a plurality of communications channels between at least one earth station and at least one user terminal through a satellite, said subsystem comprising:

   An operating center for assigning a total satellite available power for use by a satellite with at lest one communications channel, said channel being in a predetermined subband in a predetermined beam emitted by the satellite; and
   an earth station for controlling distribution of said total available power among communications channels in subbands of beams emitted by the satellite, said earth station including a baseband manager for dividing said total available power between subbands to provide subband power allocations, said earth station including a beam processor for determining a desired signal quality to be maintained by the satellite along said communications channels, said beam processor adjusting said desired signal quality based on said subband power allocations.

2. A subsystem according to claim 1, wherein said earth station further includes a plurality of modems assigned to a corresponding number of communications channels, said modems adjusting a satellite transmission power level, at which the satellite emits RF signals along of forward links of said communications channels, to maintain said desired signal quality determined by said beam processor.

3. A subsystem according to claim 1, wherein said baseband manager determines subband power allocations for each beam based on total satellite power allocation, PFD limits and required power levels reported from said beam processor.

4. A subsystem according to claim 1, wherein said beam processor determines fade margins for each user terminal within a beam based on subband power allocations.

5. A subsystem according to claim 1, wherein said beam processor determines fade margins based on

user terminal type and user terminal position within a field of view of a satellite.

6. A subsystem according to claim 1, wherein said beam processor determines fade margins based on signal-to-noise ratio information reported to said beam processor from a modem corresponding to a current user terminal.

7. A subsystem according to claim 1, wherein said baseband manager reallocates power between first and second subbands to shift power to said first subband which requires additional power and to shift power from said second subband which includes excess power.

8. A subsystem according to claim 1, wherein said beam processor decreases a desired signal-to-noise ratio for a corresponding subband when a corresponding modem indicates that sufficient power has not been allocated to said subband to achieve said desired signal-to-nose ratio.

9. A subsystem according to claim 1, wherein said beam processor reports a subband power demand for each subband to the baseband manager.

10. A subsystem according to claim 1, wherein said operations center divides a total power capacity of a common satellite between earth stations using beams of the common satellite.

11. A subsystem according to claim 1, wherein said operations center provides to said earth station power limits per subband per beam.

12. A subsystem according to claim 1, wherein said operations center reallocates a total power capacity of a common satellite between earth stations using beams of the common satellite based on feedback power requirement information from said earth stations.

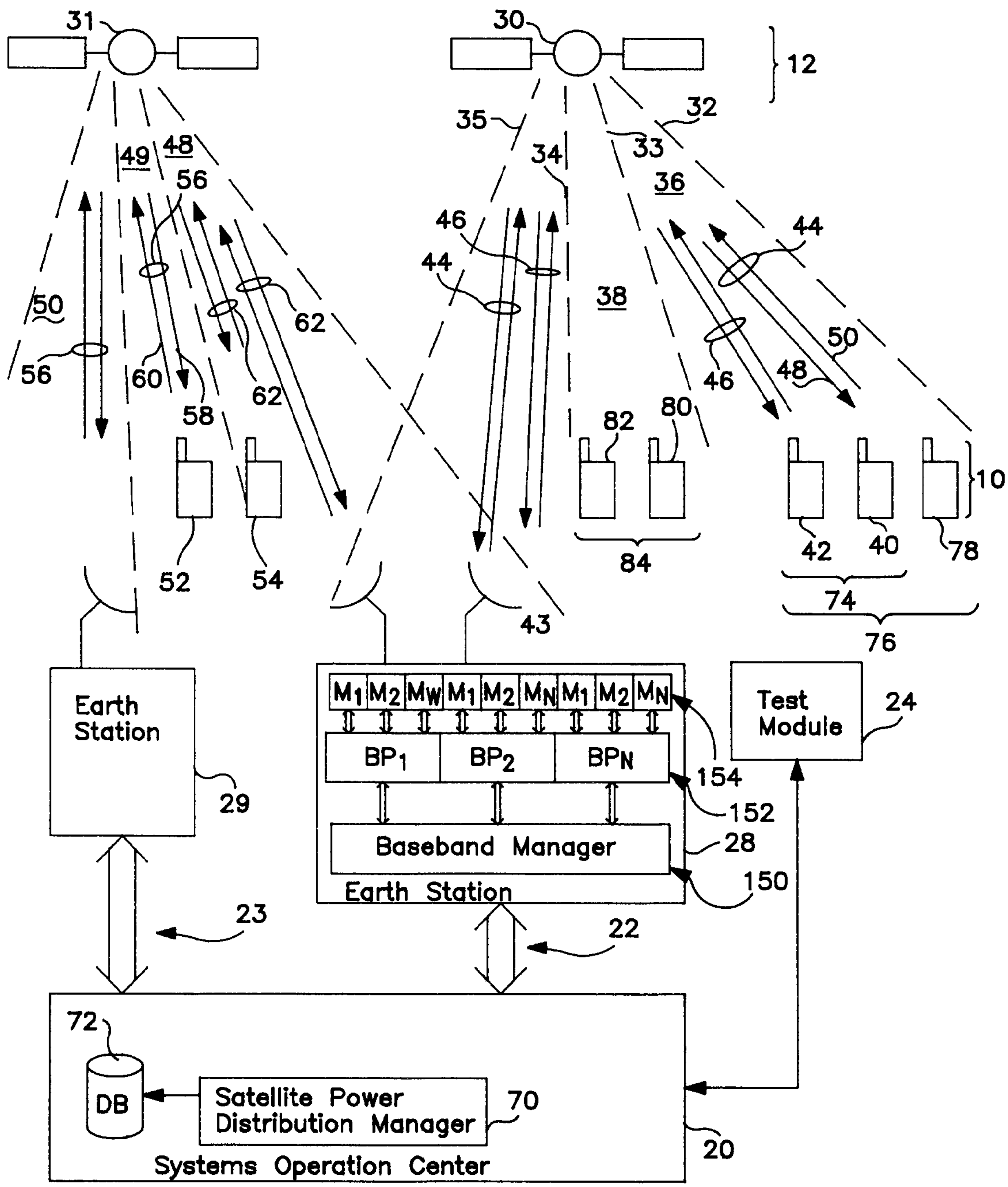

31
30
12
35
34
33
32
49
48
56
36
46
44
44
38
50
62
56
60
58
62
46
48
50
82
80
10
52
54
84
42
40
78
74
76
43
Earth
Station
29
M_1 M_2 M_W M_1 M_2 M_N M_1 M_2 M_N
BP_1 BP_2 BP_N
Baseband Manager
Earth Station
154
152
28
150
Test
Module
24
23
22
72
DB
Satellite Power
Distribution Manager
70
Systems Operation Center
20

FIG. 1

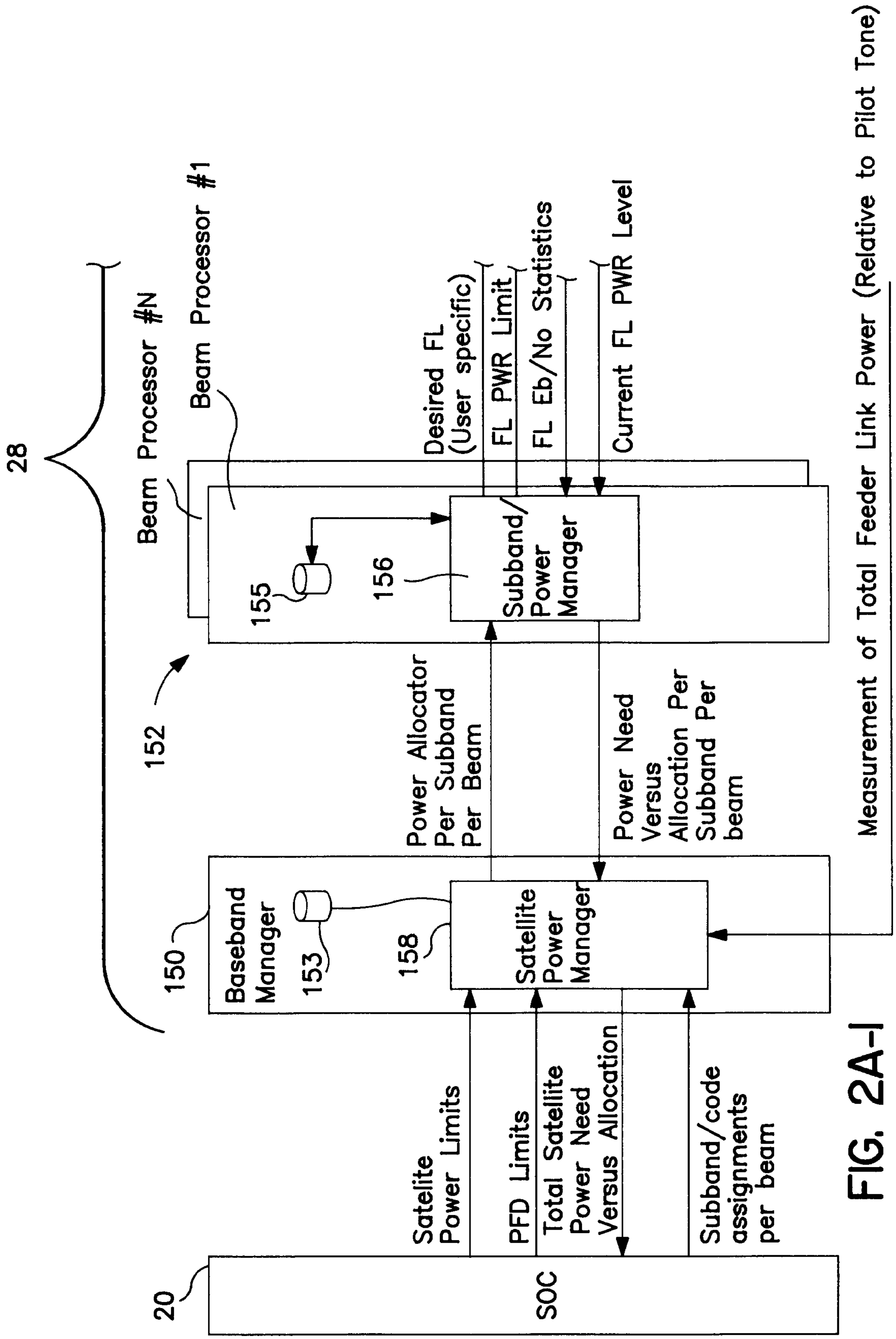

28
Beam Processor #N
Beam Processor #1
Desired FL (User specific)
FL PWR Limit
FL Eb/No Statistics
Current FL PWR Level
155
156
Subband/ Power Manager
152
Power Allocator Per Subband Per Beam
Power Need Versus Allocation Per Subband Per beam
Measurement of Total Feeder Link Power (Relative to Pilot Tone)
150
Baseband Manager
153
158
Satellite Power Manager
Satelite Power Limits
PFD Limits
Total Satellite Power Need Versus Allocation
Subband/code assignments per beam
20
SOC

FIG. 2A-I

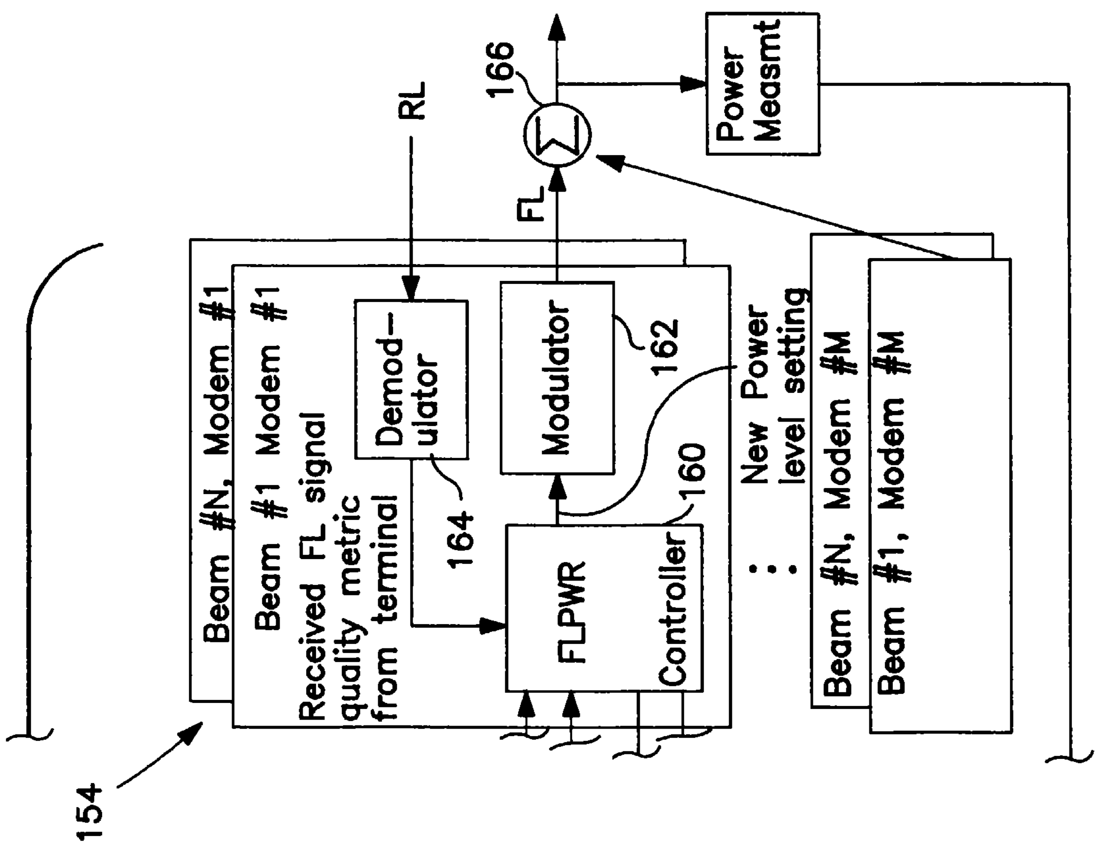
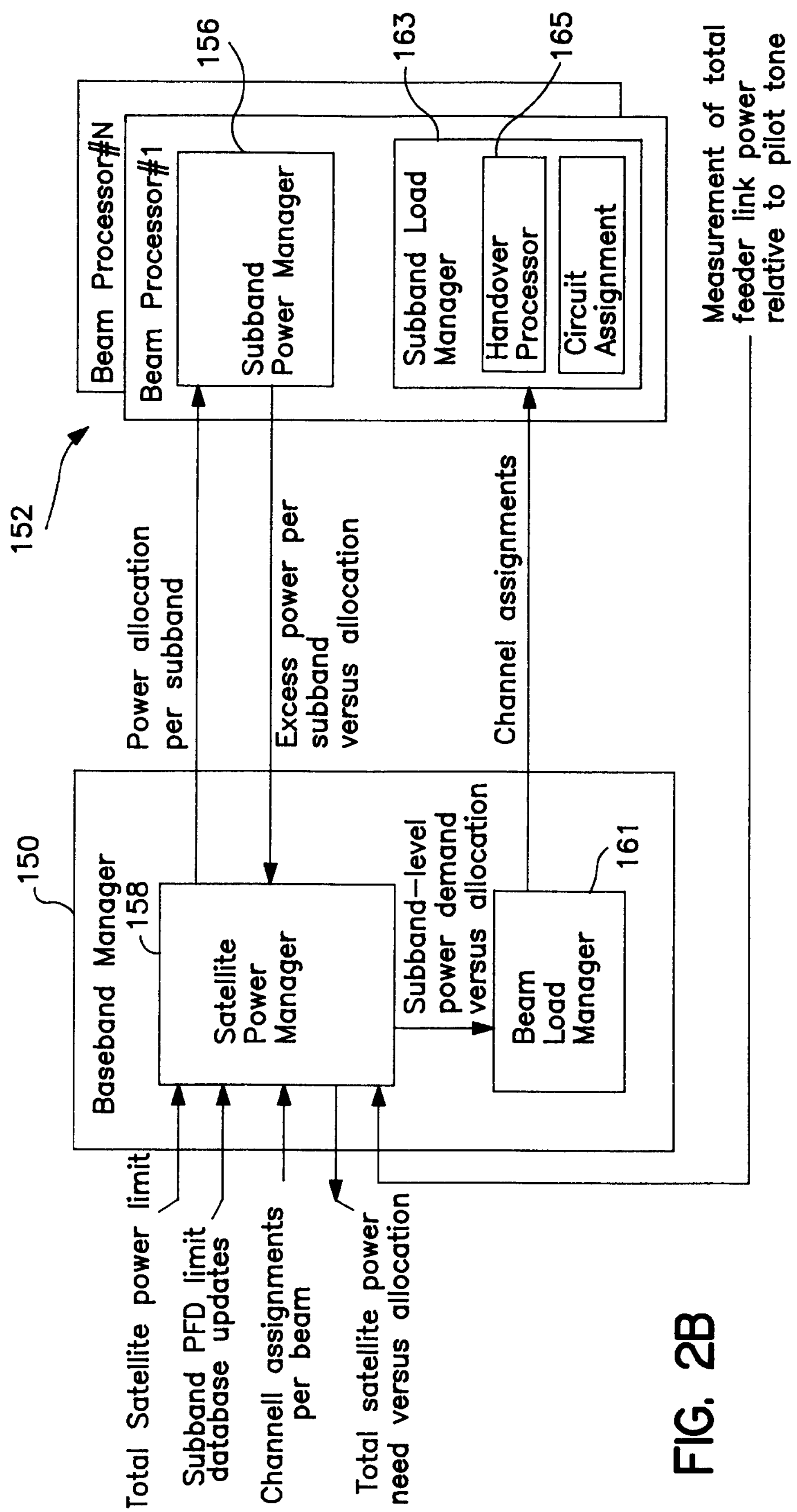

154
Beam #N, Modem #1
Beam #1 Modem #1
Received FL signal quality metric from terminal
Demod-ulator
164
Modulator
162
RL
FL
166
Power Measmt
FLPWR
Controller
160
New Power level setting
Beam #N, Modem #M
Beam #1, Modem #M

FIG. 2A-2



Beam Processor#N
Beam Processor#1
156
Subband Power Manager
163
165
Subband Load Manager
Handover Processor
Circuit Assignment
Measurement of total feeder link power relative to pilot tone
152
Power allocation per subband
Excess power per subband versus allocation
Channel assignments
150
Baseband Manager
158
Satellite Power Manager
Subband-level power demand versus allocation
161
Beam Load Manager
Total Satellite power limit
Subband PFD limit database updates
Channell assignments per beam
Total satellite power need versus allocation

FIG. 2B

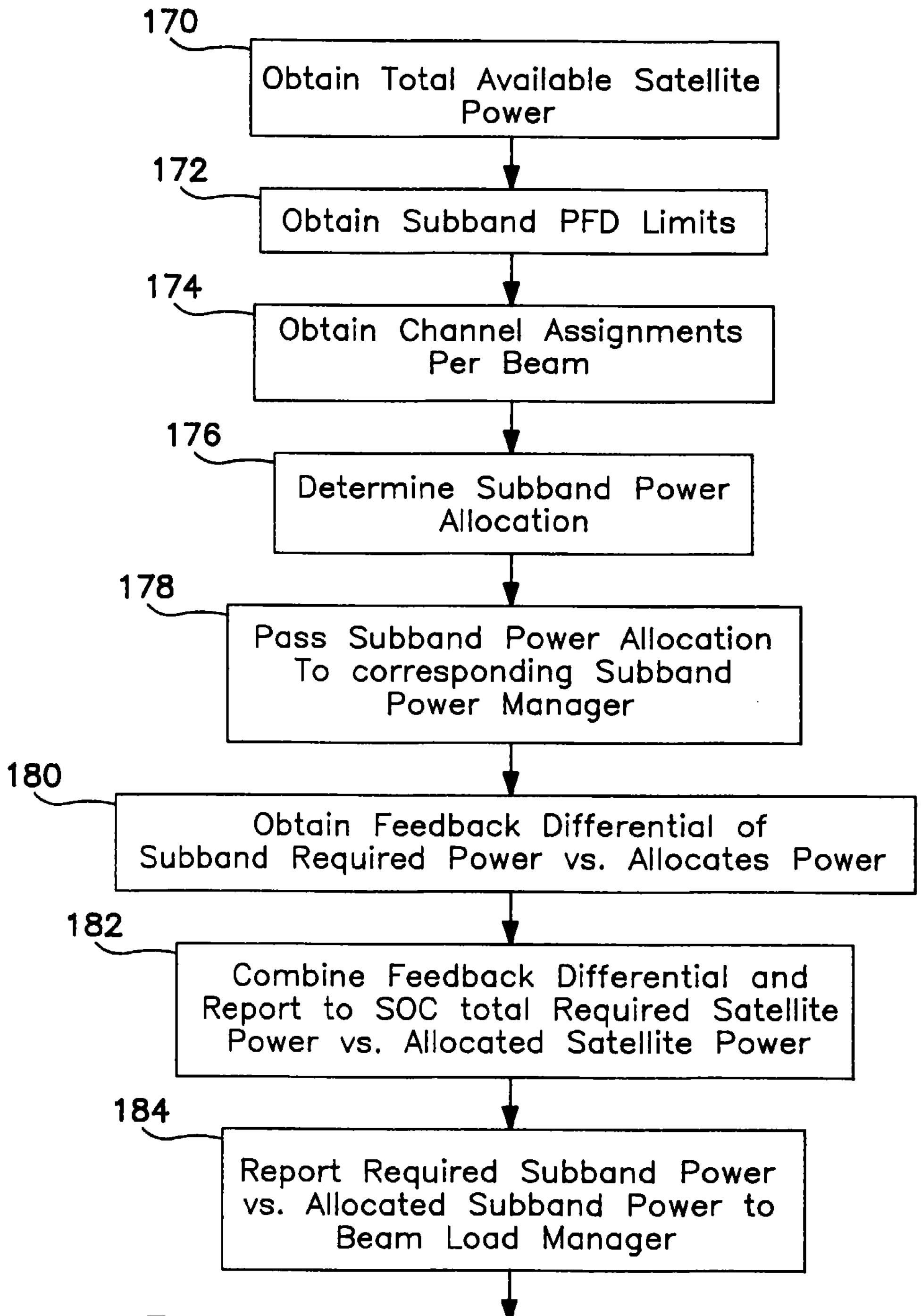

170
Obtain Total Available Satellite Power
172
Obtain Subband PFD Limits
174
Obtain Channel Assignments Per Beam
176
Determine Subband Power Allocation
178
Pass Subband Power Allocation To corresponding Subband Power Manager
180
Obtain Feedback Differential of Subband Required Power vs. Allocates Power
182
Combine Feedback Differential and Report to SOC total Required Satellite Power vs. Allocated Satellite Power
184
Report Required Subband Power vs. Allocated Subband Power to Beam Load Manager

FIG. 3A

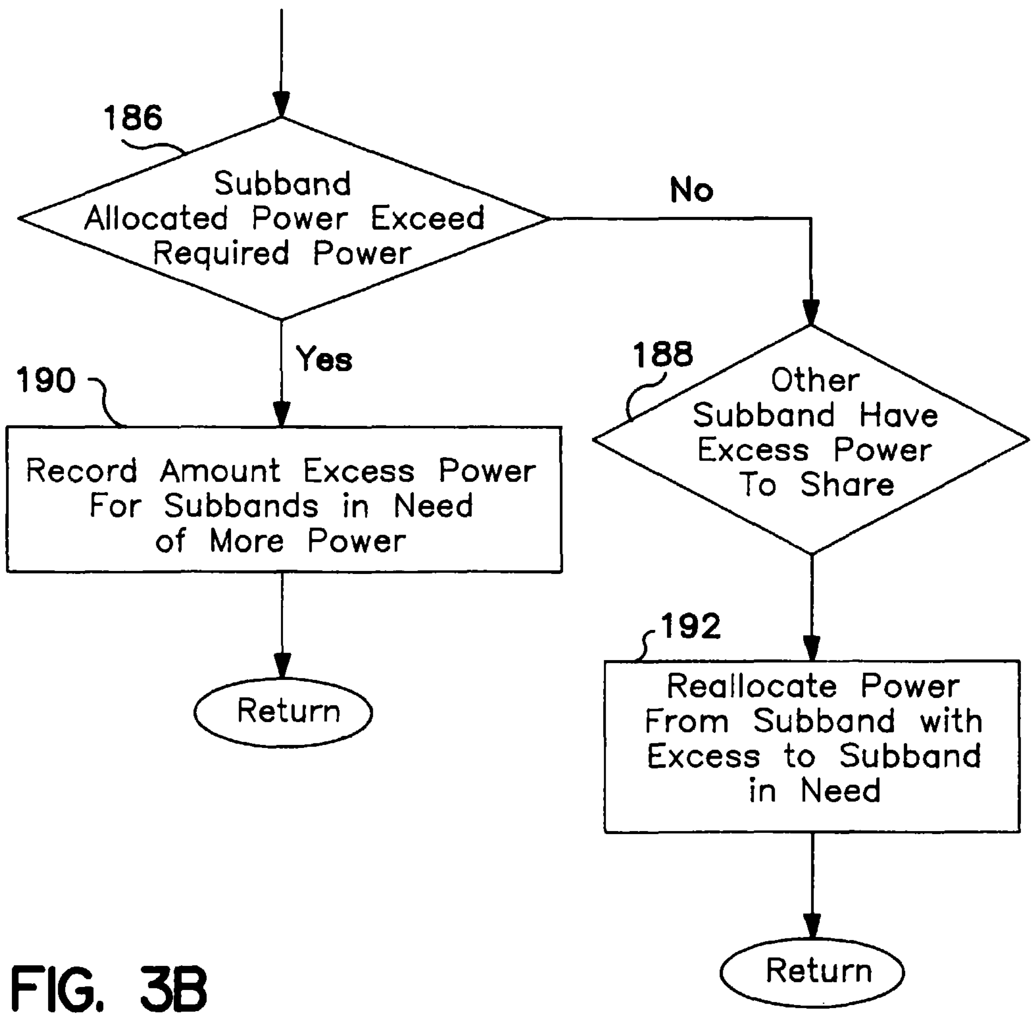

186
Subband Allocated Power Exceed Required Power
No
Yes
190
Record Amount Excess Power For Subbands in Need of More Power
Return
188
Other Subband Have Excess Power To Share
192
Reallocate Power From Subband with Excess to Subband in Need
Return

FIG. 3B

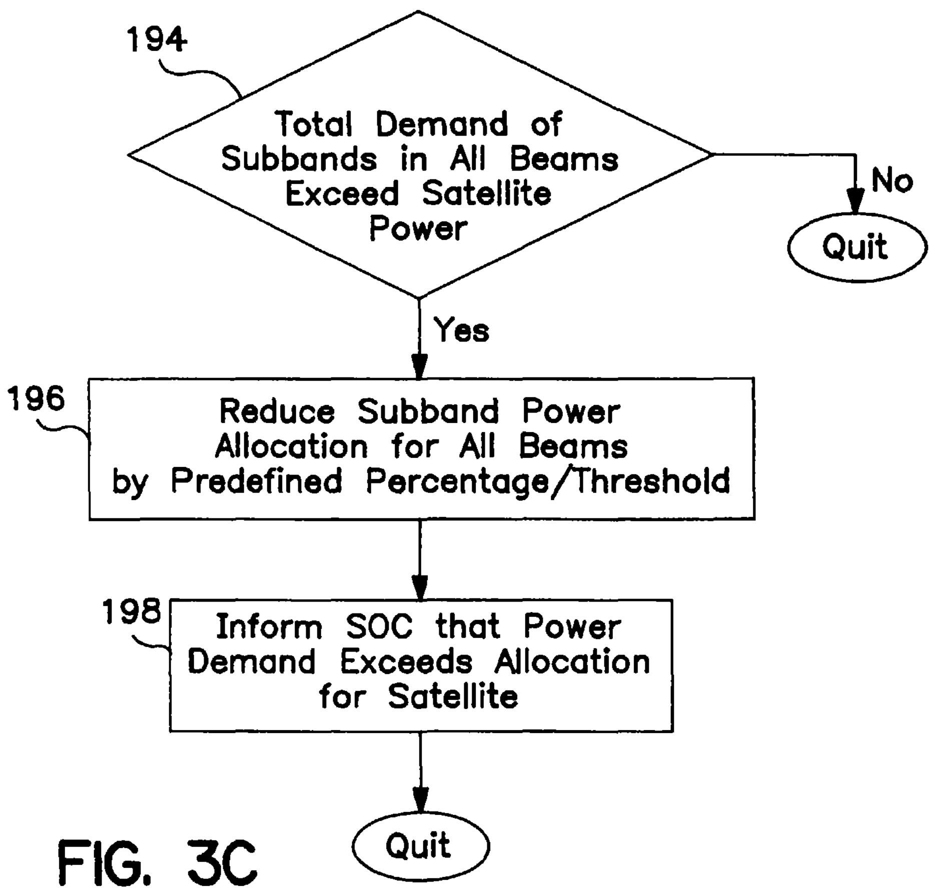

194
Total Demand of Subbands in All Beams Exceed Satellite Power
No
Quit
Yes
196
Reduce Subband Power Allocation for All Beams by Predefined Percentage/Threshold
198
Inform SOC that Power Demand Exceeds Allocation for Satellite
Quit

FIG. 3C

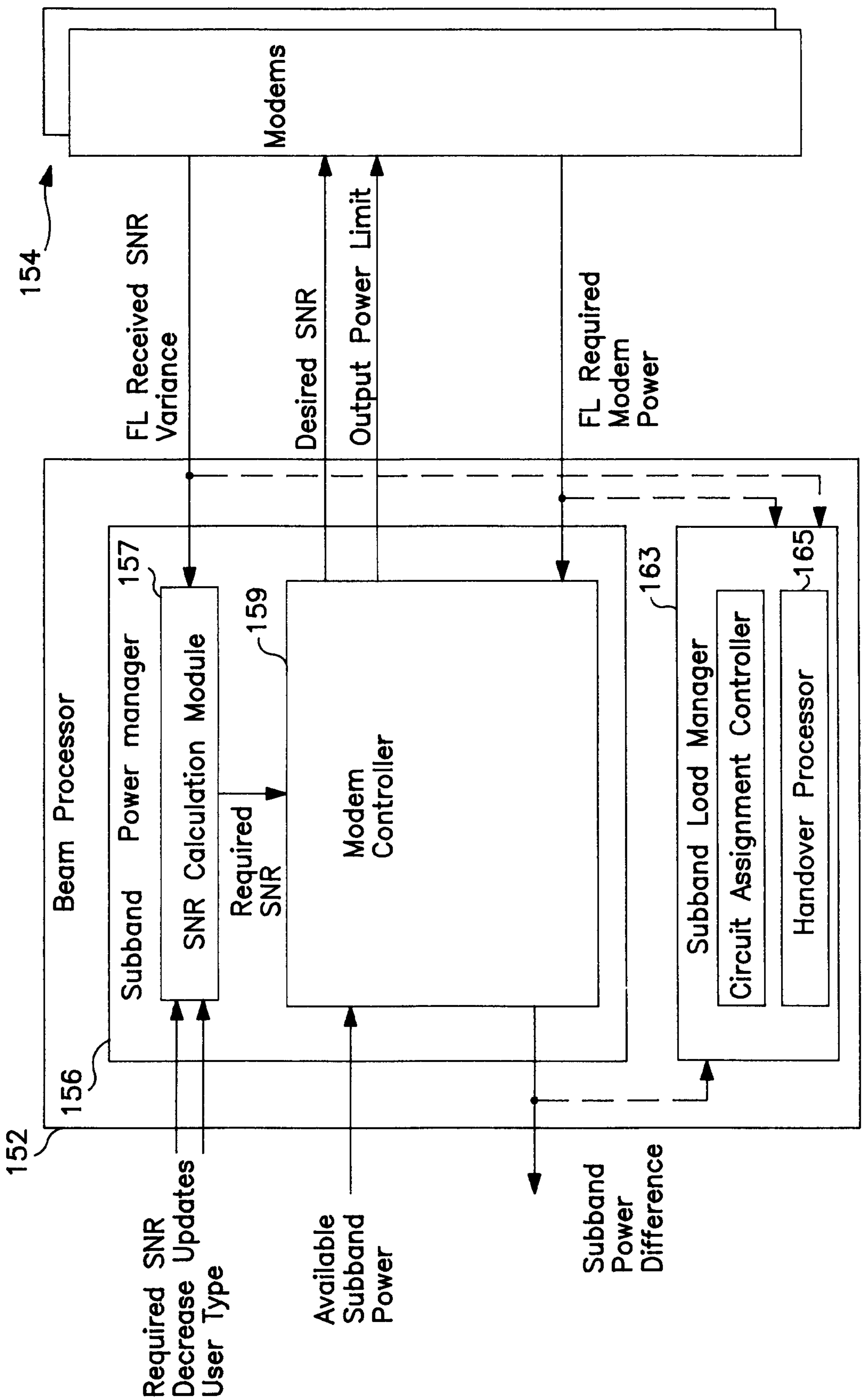

Modems
154
FL Received SNR Variance
Desired SNR
Output Power Limit
FL Required Modem Power
152
Beam Processor
156
Subband Power manager
157
SNR Calculation Module
Required SNR
159
Modem Controller
163
Subband Load Manager
Circuit Assignment Controller
165
Handover Processor
Required SNR Decrease Updates User Type
Available Subband Power
Subband Power Difference

FIG. 4

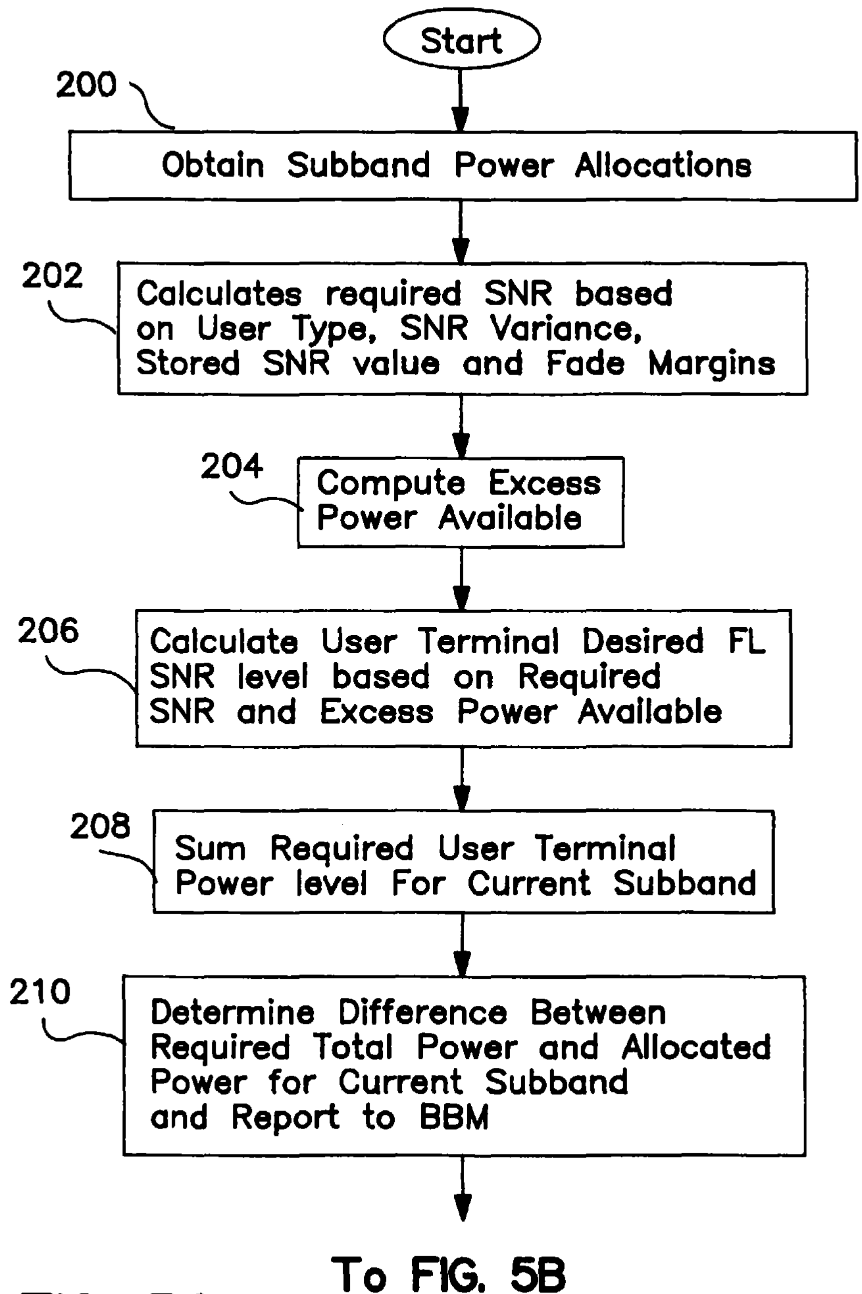

Start
200
Obtain Subband Power Allocations
202
Calculates required SNR based on User Type, SNR Variance, Stored SNR value and Fade Margins
204
Compute Excess Power Available
206
Calculate User Terminal Desired FL SNR level based on Required SNR and Excess Power Available
208
Sum Required User Terminal Power level For Current Subband
210
Determine Difference Between Required Total Power and Allocated Power for Current Subband and Report to BBM
To FIG. 5B

FIG. 5A

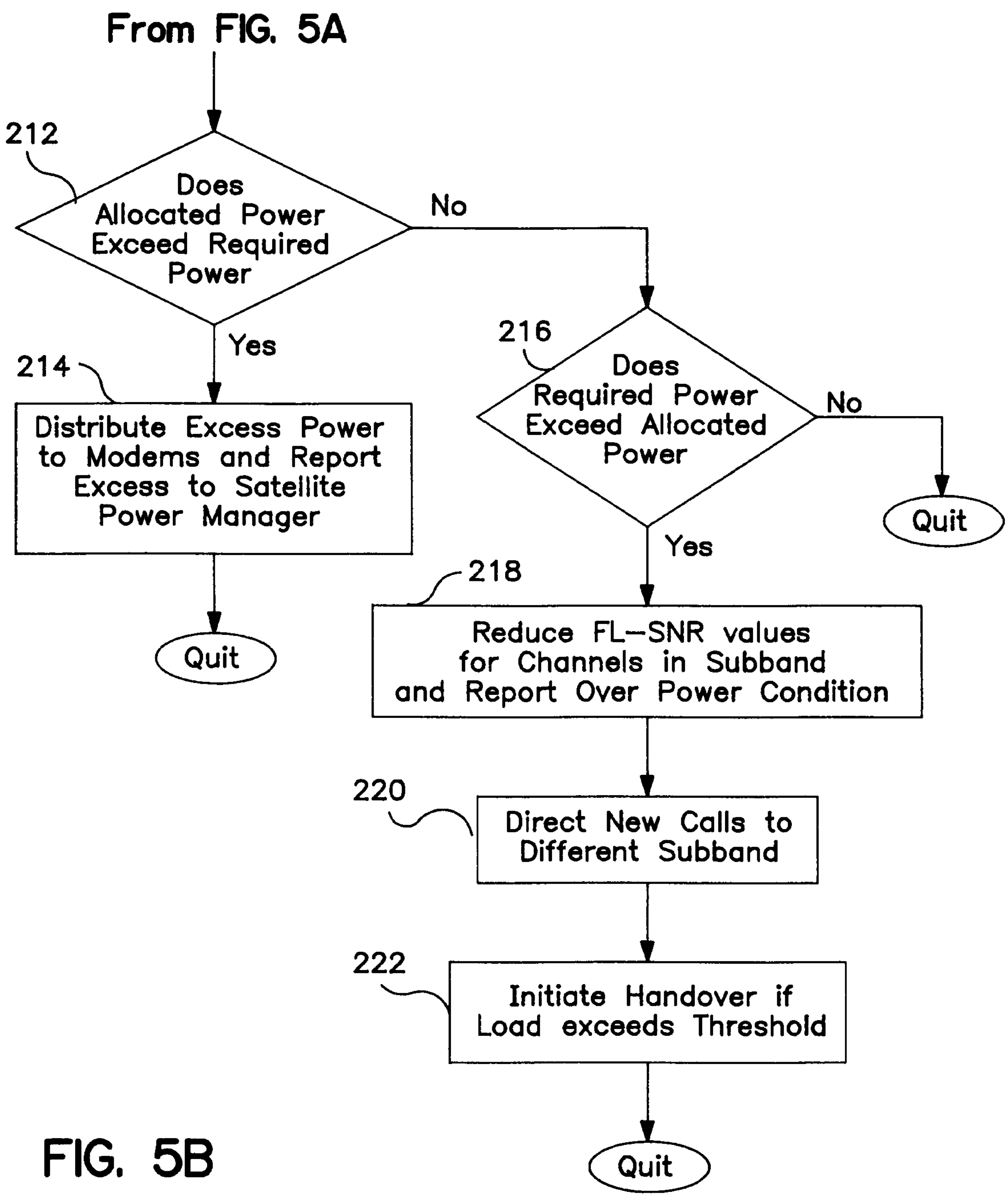

From FIG. 5A
212
Does Allocated Power Exceed Required Power
No
Yes
214
Distribute Excess Power to Modems and Report Excess to Satellite Power Manager
Quit
216
Does Required Power Exceed Allocated Power
No
Quit
Yes
218
Reduce FL-SNR values for Channels in Subband and Report Over Power Condition
220
Direct New Calls to Different Subband
222
Initiate Handover if Load exceeds Threshold
Quit

FIG. 5B

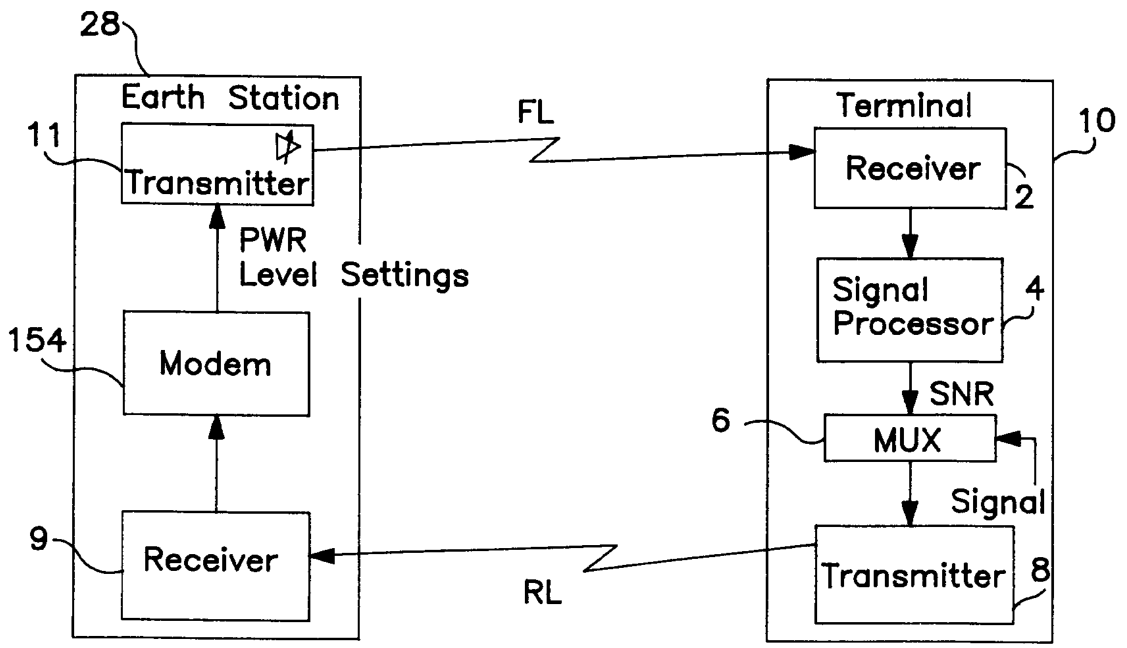

28
Earth Station
11
Transmitter
PWR
Level Settings
154
Modem
9
Receiver
FL
RL
Terminal
10
Receiver
2
Signal
Processor
4
SNR
6
MUX
Signal
Transmitter
8

FIG. 6

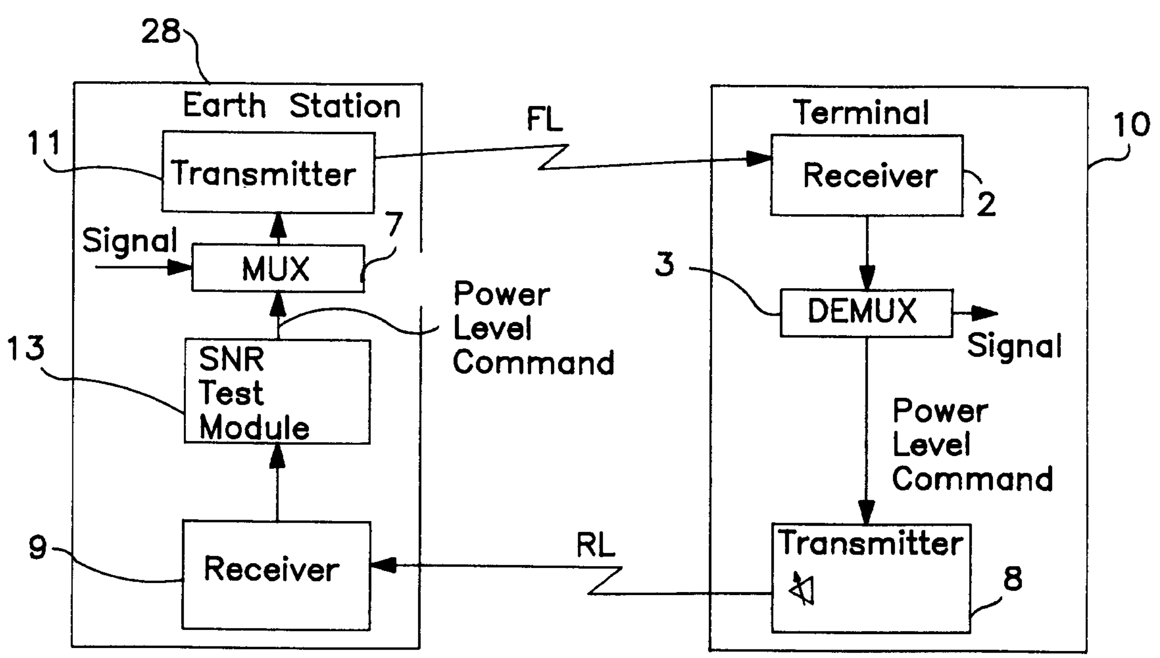

28
Earth Station
11
Transmitter
Signal
MUX
7
Power
Level
Command
13
SNR
Test
Module
9
Receiver
FL
RL
Terminal
10
Receiver
2
3
DEMUX
Signal
Power
Level
Command
Transmitter
8

FIG. 7

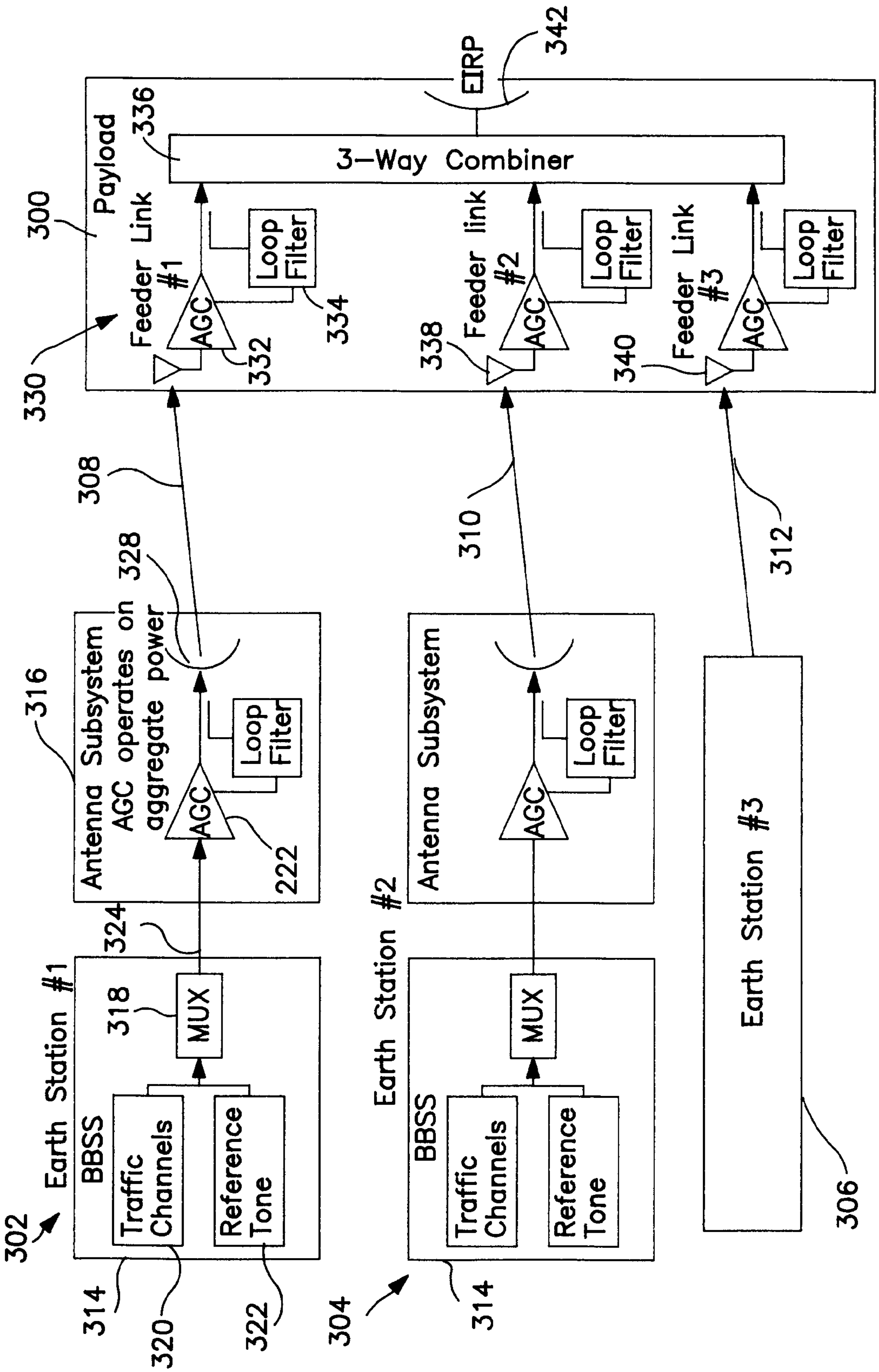

Payload
Feeder Link #1
Feeder link #2
Feeder Link #3
AGC
Loop Filter
3-Way Combiner
EIRP
Antenna Subsystem
AGC operates on aggregate power
Earth Station #1
Earth Station #2
Earth Station #3
BBSS
Traffic Channels
Reference Tone
MUX
300
302
304
306
308
310
312
314
316
318
320
322
324
328
330
332
334
336
338
340
342
222

FIG. 8

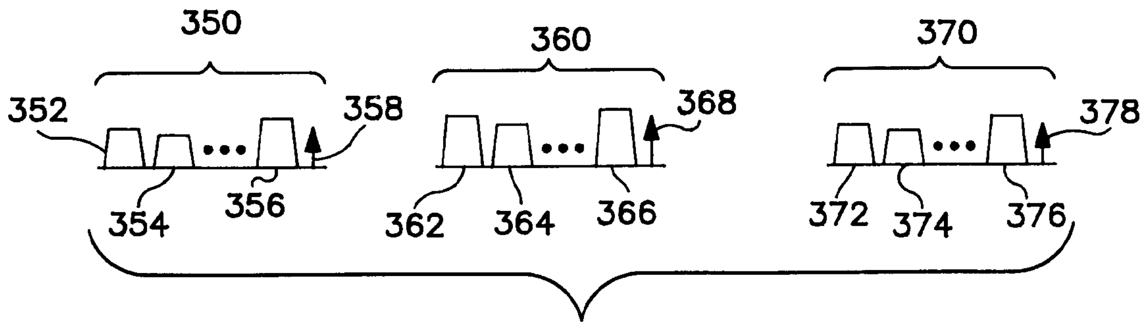

350
360
370
352
358
368
378
354
356
362
364
366
372
374
376

FIG. 9A

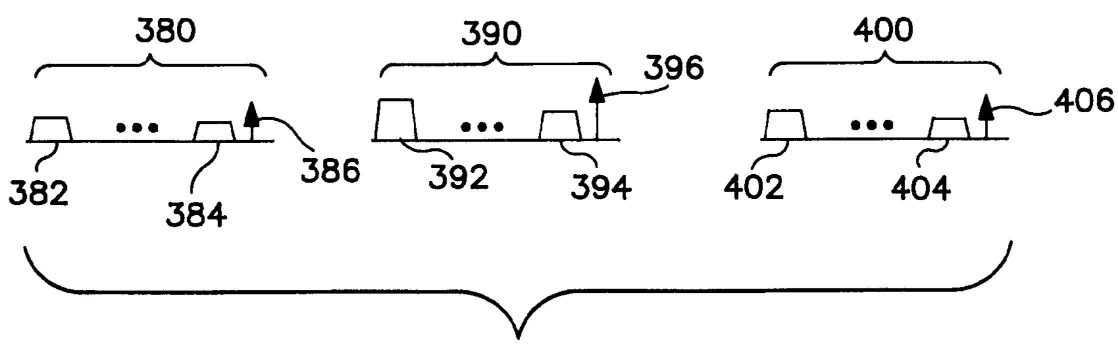

380
390
400
396
406
382
384
386
392
394
402
404

FIG. 9B

European Patent Office

EUROPEAN SEARCH REPORT

Application Number
EP 97 10 6564

DOCUMENTS CONSIDERED TO BE RELEVANT

Category	Citation of document with indication, where appropriate, of relevant passages	Relevant to claim	CLASSIFICATION OF THE APPLICATION (Int.Cl.6)
A	EP 0 275 118 A (NIPPON ELECTRIC CO) 20 July 1988 * abstract; claim 1; figures 5-7 * * column 10, line 47 - column 15, line 45 * ---	1	H04B7/185 H04B7/005
A	US 4 261 054 A (SCHARLA-NIELSEN HANS) 7 April 1981 * abstract; figure 1 * ---	1	
A	US 5 446 756 A (MALLINCKRODT ALBERT J) 29 August 1995 * abstract * -----	1	
			TECHNICAL FIELDS SEARCHED (Int.Cl.6)
			H04B

The present search report has been drawn up for all claims

Place of search	Date of completion of the search	Examiner
MUNICH	23 June 1997	Kolbe, W

CATEGORY OF CITED DOCUMENTS

X : particularly relevant if taken alone
Y : particularly relevant if combined with another document of the same category
A : technological background
O : non-written disclosure
P : intermediate document

T : theory or principle underlying the invention
E : earlier patent document, but published on, or after the filing date
D : document cited in the application
L : document cited for other reasons

& : member of the same patent family, corresponding document

Detection of the transmission mode of a DVB signal

Publication number: EP0895387 (A1)
Publication date: 1999-02-03
Inventor(s): KLANK OTTO [DE]; KLAUSBERGER WOLFGANG [DE]; LAABS JUERGEN [DE] +
Applicant(s): THOMSON BRANDT GMBH [DE] +
Classification:
- international: ***H04J11/00; H04L27/00; H04L27/22; H04L27/26; H04L27/38; H04N7/24;*** *H04H20/72;* **H04J11/00; H04L27/00; H04L27/22; H04L27/26; H04L27/38; H04N7/24;** (IPC1-7): H04H1/00; H04L27/00; H04L27/26
- European: H04L27/26M5C3; H04L27/26M5C5; H04N7/24A
Application number: EP19970112929 19970728
Priority number(s): EP19970112929 19970728

Also published as:
US6330293 (B1)
JP11074863 (A)
DE69826387 (T2)
CN1206966 (A)
CN1133280 (C)

Cited documents:
WO9726742 (A1)
DE4403408 (C1)
EP0562422 (A1)
WO9707620 (A1)
EP0653858 (A2)

Abstract of **EP 0895387 (A1)**

The reception method involves providing coarse time synchronisation, with the received digital signal (INP) correlated with time-shifted versions of itself corresponding to the possible transmission modes, for detection of the maximum correlation values, for providing the transmission mode, the protection interval and the sampling window. A coarse automatic frequency correction is provided via a multiplier (M) before Fourier transformation of the received digital signal and correlation of the Fourier transformation dependent on the identified transmission mode, with evaluation of the correlation result for determining the system conformity and reception quality.

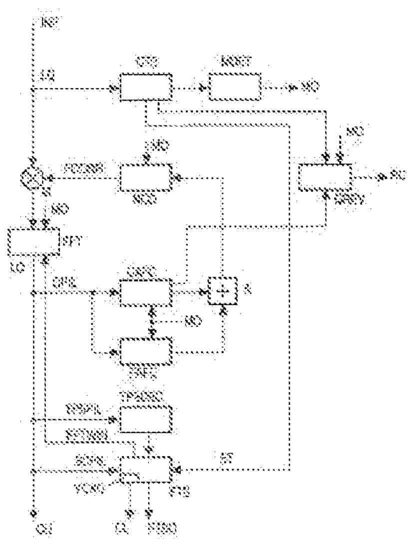


Fig.1

Data supplied from the ***espacenet*** database — Worldwide

(19) Europäisches Patentamt
European Patent Office
Office européen des brevets

(11) EP 0 895 387 A1

(12) **EUROPÄISCHE PATENTANMELDUNG**

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(51) Int. Cl.6: **H04L 27/26**, H04L 27/00, H04H 1/00

(21) Anmeldenummer: **97112929.1**

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(71) Anmelder:
DEUTSCHE THOMSON-BRANDT GMBH
78048 Villingen-Schwenningen (DE)

(72) Erfinder:
- **Klank, Otto**
  31275 Lehrte-Arpke (DE)
- **Klausberger, Wolfgang**
  30880 Laatzen (DE)
- **Laabs, Jürgen**
  30982 Pattensen (DE)

(54) **Erkennung des Übertragungsmodus eines DVB-Signales**

(57) Zur Abstimmung, beim Empfang, erfolgt eine grobe Symbolsynchronisation, wobei das Signal im Zeitbereich korreliert wird, mit verschiedenen, den möglichen Übertragungsmodi entsprechenden, zeitlichen verschobenen Kopien von sich selbst. Davon werden der aktuelle Modus, das aktuelle Schutzintervall, und ein Abtastfenster geleitet.

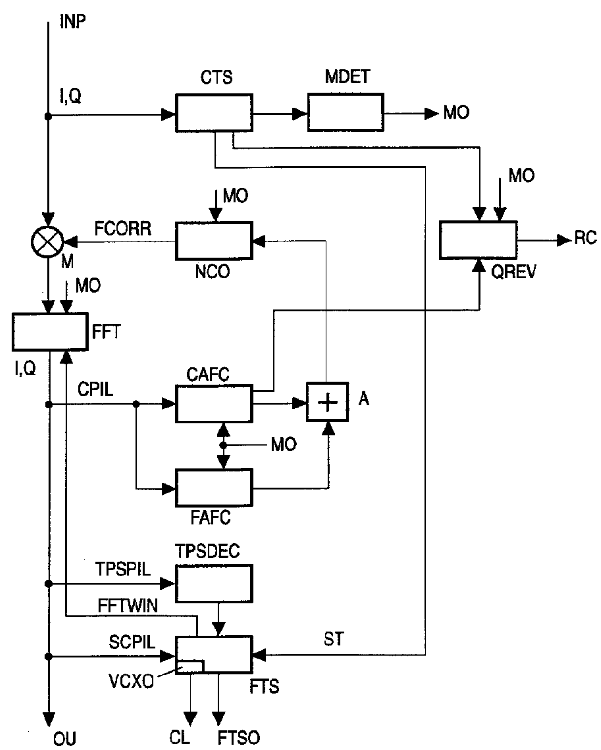

Fig.1

EP 0 895 387 A1

Beschreibung

[0001] Die Erfindung betrifft ein Verfahren und einen Empfänger für den Empfang von Multicarrier-Digitalsignalen.

Stand der Technik

[0002] Für die terrestrische Übertragung von digitalen Fernseh- und Rundfunksignalen (im folgenden allgemein Rundfunksignale genannt) können Modulation-Typen wie OFDM, QPSK und QAM verwendet werden. Beispiele für solche Rundfunksignale sind DVB (digital video broadcast), HDTV-T (hierarchical digital television transmission) und DAB (digital audio broadcast). Das DVB-System ist in seinen Grundlagen in ETS 300 744 spezifiziert.
Die Daten der digitalen Rundfunksignale sind in zweidimensionalen (Zeit und Frequenz, im Folgenden als 'temporal-spektral' bezeichnet) Rahmen oder Frames angeordnet, die eine zeitliche Länge von T_F haben und im Fall von ETS 300 744 aus 68 OFDM-Symbolen bestehen. Vier Rahmen bilden einen Super-Rahmen. Bei den obengenannten Übertragungs-Systemen für digitale Audio- oder Videosignale sind verschiedene Übertragungs-Modi möglich. Im Fall von ETS 300 744 werden die Symbole der Dauer T_s gebildet aus jeweils 1705 Trägern (2K-Mode) oder aus jeweils 6817 Trägern (8K-Mode) mit unterschiedlichen Frequenzen.
Der 2K-Mode ist insbesondere für einzelne Sender und kleine SFN-Netzwerke (single frequency network) mit begrenzten Sender-Abständen geeignet.
Der 8K-Mode kann für einzelne Sender und für kleine und große SFN-Netzwerke verwendet werden.
Die Symbole haben eine zeitliche Länge von T_s mit einem Nutzteil der Länge T_u und einem Guard-Intervall der Länge Δ. Das Guard-Intervall wird durch eine zyklische Fortsetzung des Nutzteils gebildet und ist zeitlich vor diesem angeordnet. Alle Symbole enthalten Daten und Referenz-Information. Jedes Symbol kann als eine Gruppe von Zellen angesehen werden, wobei jedem Träger eine Zelle entspricht.
Die Rahmen enthalten außer den eigentlichen Bild-, Ton-, oder sonstigen Daten verstreute Pilotzellen (scattered pilots), kontinuierliche Pilotsignalen und TPS-Träger oder -Pilots (transmission parameter signalling) . Diese sind z.B. in ETS 300 744, März 1997, in den Kapiteln 4.4. bis 4.6 beschrieben.
Die Pilotzellen bzw. -Träger enthalten Referenz-Information, deren übertragener Wert dem Empfänger bekannt ist. Die kontinuierlichen Pilotsignalen können beispielsweise bei jedem vierten Symbol mit einer verstreuten Pilotzelle in Koinzidenz sein. Der Wert bzw. Inhalt der verstreuten und kontinuierlichen Pilotsignalen ist z.B. von einer Pseudo-Zufalls-Binärsequenz w_k für jeden der übertragenen Träger k abgeleitet. Die Sequenz w_k kann auch die Startphase der TPS-Trägerinformation bestimmen. Die Pilotzellen bzw. -Träger können empfängerseitig zur Rahmen-Synchronisation, Frequenz-Synchronisation, Zeit-Synchronisation, Kanalschätzung und Übertragungs-Mode-Identifizierung verwendet werden. Ob und wie diese Möglichkeiten empfängerseitig genutzt werden, bleibt dem Empfänger-Hersteller überlassen.
In EP-A-0 786 889 ist ein entsprechendes System für die Ahwendung bei DAB beschrieben.

Erfindung

[0003] Ein wichtige Überlegung im Zusammenhang mit solchen Systemen ist es, ein systemkonformes Signal in dem Fall zu finden, wo ein Empfänger eingeschaltet oder auf einen anderen Kanal abgestimmt wird. Dazu muß der Empfänger verschiedenen Dienste voneinander unterscheiden können, z.B. digitale Signale von analogen Signalen oder digitale DVB-Signale von digitalen DAB-Signalen. In bestimmten Frequenzbereichen können sowohl Digitalsignale als auch Analogsignale (z.B. PAL-Fernsehsignale) vorkommen, wobei die Zentrums-Frequenzen von den vorgegebenen Kanal-Mittenfrequenzen abweichen können.
[0004] Der Erfindung liegt die Aufgabe zugrunde, ein verbessertes Verfahren zur Abstimmung beim Empfang von Multicarrier-Digitalsignalen oder zur Überprüfung der Systemkonformität solcher empfangenen Digitalsignale anzugeben. Diese Aufgabe wird durch das in Anspruch 1 angegebene Verfahren gelöst.
[0005] Der Erfindung liegt die weitere Aufgabe zugrunde, einen Empfänger zur Anwendung des erfindungsgemäßen Verfahrens anzugeben. Diese Aufgabe wird durch den in Anspruch 13 angegebenen Empfänger gelöst.
[0006] Empfängerseitig wird zunächst eine grobe Zeit-Synchronisation verbunden mit einer Mode-Detektion und eventuell zusätzlich eine grobe AFC (automatic frequency correction) sowohl beim Suchen und Identifizieren von Empfangssignalen als auch bei deren ständiger Überwachung durchgeführt.
Bei der groben Zeit-Synchronisation wird das Zeitsignal mit dem um eine Nutzsymbollänge T_u verschobenen Zeitsignal korreliert. Diese Korrelation kann mehrfach, z.B. fünfmal pro Daten-Frame durchgeführt werden. Bei dieser Korrelation werden Signalproben mit verschiedener Länge T_u entsprechend dem jeweiligen Mode verwendet und anhand der sich ergebenden Korrelations-Ergebnis-Maxima wird dann auf den aktuell vorliegenden Mode (z.B. 2K- oder 8K-Mode) geschlossen. Falls sich kein brauchbares Korrelations-Ergebnis-Maximum ergibt, können die Korrelationsschritte wiederholt werden.
Aus dem Abstand und/oder der Amplitude der Maxima wird unter Berücksichtigung des Modes das verwendete Guar-

dintervall ermittelt und nachfolgend ein Abtastfenster positioniert.
Dies kann durch einmaliges Setzen eines zur Symbolfolge ($T_u+\Delta$) synchronen Zählers erfolgen, der ein Zeitfenster mit Dauer T_u ausgibt. Dieses Zeitfenster wird im folgenden auch mit Abtastfenster oder FFT-Fenster bezeichnet. Ein dabei benutzter Basisoszillator und damit auch die Position des Fensters werden in späteren Schritten über eine Zeit-Feinsynchronisation nachgeregelt.

[0007] Falls der Mode richtig erkannt und das Abtastfenster annähernd korrekt positioniert wurde, kann eine FFT mit dem Mode entsprechender Länge erfolgen. Statt einer FFT kann bei der Erfindung ganz allgemein eine Fourier-Transformation oder eine sonstige Transformation, die eine frequenz-spektrale Darstellung vom Zeitbereich in den Frequenzbereich ermöglicht, zum Einsatz kommen. Aus dem so umgewandelten Signal werden Pilotzellen nach dem vorgesehenen Anordnungsschema entnommen und mit dem gemäß Spezifikation vorgesehenen Werten korreliert. Nach Spezifikation sind z.B. beim 2K-Mode 45 Positionen und beim 8K-Mode 177 Positionen des Spektrums mit kontinuierlichen Pilotsignalen belegt. Zur Korrelation werden z.B. beim 2K-Mode ±16 solcher Sätze (über ±16 Trägerabstände) und beim 8K- Mode ±64 solcher Sätze (über ±64 Trägerabstände) benutzt. Über die durchgeführten Korrelationsschritte ergibt sich ein Korrelationsmaximum und möglicherweise in direkter Nachbarschaft einige Nebenmaxima mit geringerer Amplitude. Aus der Position des Maximums kann die Frequenzablage des Basisband-Signals ermittelt werden. Dieses Ergebnis wird zur groben Korrektur der Frequenz, z.B. mittels eines vor dem FFT-Teil angeordneten Multiplizieres M benutzt, so daß bei weiteren Schritten die Frequenz-Abweichung kleiner ist als ±1/2 Trägerabstand.

[0008] Voraussetzung ist jedoch, daß die Position des Maximums vorher mit genügender Sicherheit bzw. einer Genauigkeit von besser als ±1/2 Trägerabstand erkannt wurde. Zur genaueren Abschätzung der Position $l_{real,s}$ des Maximums kann folgende Berechnung durchgeführt werden:

$$l_{real,s} = l_{max,s} + W_{lmax,s,1} / (W_{lmax,s} + W_{lmax,s,1}) * \operatorname{sgn}(l_{max,s,1} - l_{max,s}),$$

wobei "sgn" das Vorzeichen der Positions-Differenz ist und der Wert des größten Maximums $W_{lmax,s}$ ist und an Position $l_{max,s}$ liegt und der nächstkleinere, mit $W_{lmax,s,1}$ bezeichnete Maximum-Wert (mit gleicher Polarität) entweder an der als $l_{max,s,1}$ bezeichneten Position $l_{max,s\,+1}$ oder $l_{max,s\,-1}$ liegt.

[0009] Diese Berechnungen können dadurch vereinfacht werden, daß die beiden Werte - das Maximum und das nächstkleinere Maximum - in der Reihenfolge der l-Werte benutzt werden. Die möglichen Positionen werden dann als $l_{1,s}$ (die erste Position) und $l_{2,s}$ bezeichnet, wobei das Maximum entweder bei $l_{1,s}$ oder bei $l_{2,s}$ liegen kann. Die Vorzeichen-Funktion verschwindet dann:

$$l_{real,s} = l_{1,s} + W_{l2,s} / (W_{l1,s} + W_{l2,s}) .$$

[0010] Zur Verbesserung der AFC können mehrere, vorzugsweise drei, solcher (zeitlich nacheinander gewonnener) Ergebnisse kombiniert oder gefiltert bzw. gemeinsam verarbeitet werden. Die nächste Frequenz-Auswertung kann nach größerem Abstand erfolgen, z.B. können zum Zweck der Synchronisations-Kontrolle insgesamt 3 bis 6 solcher Auswertungen pro Frame durchgeführt werden, um den Rechenaufwand in vernünftigen Größenordnungen zu halten.

[0011] Der auf diese Weise ermittelte Zwischenwert bzw. genauere Wert der Abweichung wird bereits bei der oben beschriebenen Frequenzkorrektur berücksichtigt. Die Frequenz-Grobeinstellung mit einer besseren Genauigkeit als ± 1/2 Trägerabstand ($-F_s/2 < \Delta f < F_s/2$) ist die Voraussetzung für die folgende Übernahme der AFG-Funktion durch die sogenannte Feinregelung.
Die erreichte Genauigkeit kann nach durchgeführter Grobeinstellung durch nochmaliges Überprüfen der Frequenz ermittelt werden. In diesem Fall sollte das Ergebnis $-F_s/3 < \Delta f < F_s/3$ sein. Wird dies nicht erreicht bzw. führt die anschließende Feinkorrektur zu einem Zustand, bei dem eine weitere Signalverarbeitung (Decodierung) nicht möglich ist, so müssen die oben beschriebenen Vorgänge mit einer um einen Trägerabstand versetzten Position (in Richtung der Seite mit dem nächstniedrigen oder möglicherweise gleichgroßen Korrelations-Ergebnis wiederholt werden.

[0012] Nach der groben Zeit-Synchronisation und/oder der groben AFC erfolgen bestimmte Auswertungen. Sowohl mit den Werten aus der zeitlichen Korrelation als auch mit denjenigen der Korrelation über der Frequenz wird (jeweils) ein Verhältnis gebildet aus dem ermittelten Wert des Maximums (bzw. Zentrumswert bei der zeitlichen Korrelation) und dem Durchschnittswert der nicht zum Maximum bzw. zum Zentrumsbereich gehörenden übrigen Korrelations-Teilergebnisse.
Aus den Ergebnissen der zeitlichen Korrelation kann z.B. ein Bereich der Länge T_u herausgegriffen werden, wobei das

Maximum nicht unbedingt in der Mitte liegen muß. Ein Bereich mit einer Länge von ±1/2 Guardintervall-Länge ist bei der Durchschnittswert-Berechnung auszusparen. Das Zentrum des Zentrumsbereiches kann z.B. dadurch ermittelt werden, daß die Punkte bei -6dB ermittelt werden und eine Mittelposition berechnet wird. Dies reduziert vorteilhaft den Einfluß von Rauschen und von Multipath-Effekten.
Für die Auswertung der über der Frequenz ermittelten Korrelations-Teilergebnisse wird z.B. der gesamte Bereich von ±16 Einzelschritten (beim 2K-Mode) bzw. ±64 Einzelschritten (beim 8K-Mode) benutzt. Auch hierbei kann das Hauptmaximum außerhalb der Mitte liegen und es können Nebenmaxima mit größerer Distanz vorhanden sein. Im Gebiet von $\pm F_s$ um das Hauptmaximum können ebenfalls Nebenlinien bestehen, die jedoch zum Hauptmaximum zu rechnen sind und durch eine Abweichung der Signallage vom Raster F_s in der Größe von $-F_s/2 \leq \Delta f \leq F_s/2$ entstehen. Für die Auswertung empfiehlt es sich daher, den Maximalwert des Hauptmaximums und den benachbarten nächstkleineren Wert zusammenzufassen.
Der Durchschnittswert C_{av} wird z.B. als quadratischer Mittelwert aller nicht zum Hauptmaximum bzw. Zentrumsbereich gehörenden Korrelations-Teilergebnisse berechnet:

$$C_{av} = \sqrt{\left(\left(\sum_{0}^{l_1} |W_l|^2 + \sum_{l_2}^{l_{max}} |W_l|^2 \right) / \left(l_1 + l_{max} - l_2 + 2 \right) \right)},$$

wobei der Bereich l_1+1 bis l_2-1 den ausgesparten Teil betrifft. Bei komplexen Teilergebnissen (W_l) kann anstelle der Betragsbildung auch die Summe der Quadrate der Real- und Imaginärteile gebildet werden. Unter praktischen Gesichtspunkten sind weitere Vereinfachungen möglich, z.B. können bei Umstellung der Formel und entsprechend geänderten Mindestwerten anstelle der Division und der Berechnung der Quadratwurzel Multiplikationen durchgeführt werden, d.h. Quadrierung der Maximalwerte und Multiplikation mit dem in der Formel benutzten Divisor. Unter günstigen Signalbedingungen sowie geringeren Anforderungen an die Qualität der Aussage kann es ausreichen, nur den einfachen Mittelwert zu berechnen. Es ist ferner möglich, die Korrelations-Teilergebnisse mit einem vom Maximalwert (bzw. von der Summe aus dem Maximalwert und dem benachbarten nächstkleineren Wert) abgeleiteten Grenzwert einzeln zu vergleichen und daraus Aussagen über das Korrelationsergebnis insgesamt abzuleiten. Dies ist dann möglich, wenn generell ein ausreichender Abstand des Maximums zu den übrigen Teilergebnissen gewährleistet werden kann, was bei der Korrelation über der Frequenz der Fall ist.
Anschließend wird geprüft, ob das aus der zeitlichen Korrelation abgeleitete (erste) Verhältnis einen vorher festgelegten ersten Mindestwert überschreitet und das aus der über der Frequenz durchgeführten Korrelation abgeleitete (zweite) Verhältnis einen vorher festgelegten zweiten Mindestwert überschreitet. Überschreitet zumindest das erste Verhältnis den Mindestwert, oder optional, überschreiten beide Verhältnisse die für sie festgelegten Mindestwerte, so wird das empfangene Signal als systemkonform angesehen. Ist zumindest eine der Bedingungen nicht erfüllt, so wird das Signal als nicht-systemkonform angesehen.

[0013] Abhängig vom Ergebnis wird ein Empfangssignal während des Suchens bzw. beim Versuch ein bestimmtes Signal zu empfangen oder beim laufenden Empfang als "systemkonform" bzw. "vorhanden" oder "nicht-systemkonform" bzw. "nicht vorhanden" gekennzeichnet.

[0014] Die durchgeführten Prüfungen ergeben eine hohe Aussage-Sicherheit, bzw. die Wahrscheinlichkeit einer Falsch-Aussage ist extrem gering. Dies bedeutet, daß die nächsten Schritte gezielt erfolgen können. Bei negativem Ergebnis (d.h. keine System-Konformität) ist es z.B. nicht mehr notwendig, eine Decodierung des Signals einzuleiten, um die Konformität erneut zu überprüfen. Bei Signal-Suchvorgängen kann hierdurch viel Zeit eingespart und somit eine unnötige Wartezeit für den Benutzer des Empfängers vermieden werden.
Je nach aktuellem Kennzeichnungs-Zustand wird also beim Suchvorgang oder bei einer Empfangsprobe entweder die weitere Decodierung des Signals eingeleitet oder der Suchvorgang fortgesetzt oder bei der Empfangsprobe die Information "nicht vorhanden" ausgegeben.

[0015] Falls der Abstimmvorgang aufgrund der oben beschriebenen Ergebnisse weitergeführt werden soll, kann nun die Fein-AFC erfolgen. Dazu werden z.B. laufend die Phasenänderungen der kontinuierlichen Pilotsignale zwischen jeweils zwei aufeinanderfolgenden Symbolen einzeln bestimmt, die Ergebnisse gemittelt, aus den so ermittelten Endergebnis eine Frequenz-Abweichung berechnet und mit dieser eine Frequenzkorrektur des Signals vor der FFT durchgeführt. Die symbolweise nacheinander ermittelten Endergebnisse bzw. Frequenzabweichungen können vorteilhaft noch über mehrere Symbole zusammengefaßt und gefiltert werden.

[0016] Anschließend kann eine Rahmen-Synchronisation und eine Zeit-Fein-Synchronisation bzw. Abtast-Takt-Justierung erfolgen. Dies geschieht z.B. durch eine zeitliche Auswertung (pulse response) der 'scattered Pilots' und entsprechende Nachregelung des Abtast-Takt-Referenzoszillators, wobei (wiederum) zweckmäßigerweise mehrere zeitlich aufeinanderfolgende Werte zusammengefaßt und gefiltert werden.

[0017] Auch während des normalen Empfangs ist es zweckmäßig, die Überprüfung der groben Zeit - und Frequenz-Synchronisation (wie oben beschrieben) in gewissen Abständen vorzunehmen. Damit ist es möglich, einen Signalausfall oder ein Verschlechtern der Empfangsbedingungen oder einen Synchronisationsverlust des Empfängers schnell zu detektieren. Die Bedingungen dafür sind, daß Δt und Δf die Grenzwerte überschreiten oder die berechneten Verhältnisse die Mindestwerte unterschreiten. Der Ausdruck Δt bezeichnet dabei die Abweichung des Zentrums der Impuls-Antwort (pulse response) von der Soll-Position.
Notwendige Gegen-Maßnahmen können schnell eingeleitet werden. Würde man das Erkennen eines solchen Zustands aus den Decodierprozessen ableiten (z.B. durch starken Anstieg der Fehlerrate), so wäre unter Umständen schon sehr viel Zeit verloren.
Bei der Synchronisations-Kontrolle bzw. der laufenden Überwachung des Signals oder des Empfangs wird in dem Fall, wenn der Kennzeichnungs-Zustand auf "nicht-systemkonform" übergeht, ein Kontroll- oder Warnsignal an die übrigen Teile des Empfängers ausgegeben, so daß unter bestimmten Bedingungen - z.B. Ausfall einiger Symbole - entsprechende Maßnahmen eingeleitet werden können wie z.B. ein "Einfrieren" des letzten akzeptablen Bildes und/oder ein Muting des Tonkanals.

[0018] Vorteilhaft können zum Erkennen bzw. zur Kennzeichnung des Signalzustandes während des laufenden Betriebs noch weitere Zustandsmeldungen, wie z.B. ständig gesetztes Fehler-Flag des Viterbi-Decoders, mitausgewertet werden.

[0019] Ein Vorteil der Erfindung liegt darin, daß die Sicherheit bei der Signal-Identifizierung wesentlich erhöht wird und die Identifizierung an frühestmöglicher Stelle innerhalb der empfangsseitigen Signaldecodierung und damit auch zum frühestmöglichen Zeitpunkt erfolgt, so daß keine unnötigen Wiedergabe-Unterbrechungen eingeleitet werden müssen. Andererseits erfolgt aber eine unbedingt notwendige Unterbrechung schnell. Damit können unvertretbare Störungen wie z.B. der Ausfall oder die falsche Decodierung von einigen oder sogar allen Bildpunkt-Blöcken eines Bildes bzw. laute oder abrupte Störgeräusche beim Ton weitestgehend vermieden werden.

[0020] Im Prinzip besteht das erfindungsgemäße Verfahren darin, daß für den Empfang von Multicarrier-Digitalsignalen, die in temporal-spektralen Rahmen angeordnet sind und Daten-Symbole mit einem Guardintervall und einer Nutzsymbollänge T_u und Referenzinformationen enthalten, und die in verschiedenartigen Modi übertragen werden können, folgende Schritte zur Abstimmung beim Empfang oder zur Überprüfung der Systemkonformität der empfangenen Signale durchgeführt werden:

- grobe Zeitsynchronisation, bei der das Digitalsignal in zeitlicher Richtung korreliert wird mit dem um verschiedene, den möglichen Modi entsprechende Werte von T_u zeitlich verschobenen Digitalsignal, wobei der aktuelle Modus aus der Lage und den Beträgen von Maxima der Korrelationswerte ermittelt wird und das aktuelle Guardintervall aus Abständen von Maxima der Korrelationswerte ermittelt wird und danach ein daraus sich ergebendes Abtastfenster mit einer T_u entsprechenden Länge für Transformationsmittel und eine sich anschließende Signalauswertung gesetzt wird;
- grobe AFC-Korrektur mit Hilfe von vor den Transformationsmitteln angeordneten Multipliziermitteln und mit Hilfe von nach den Transformationsmitteln angeordneten Grob-AFC-Mitteln, wobei dem Anordnungs-Schema der Referenzinformationen entsprechende Informationen des aktuellen Symbols dem Ausgangssignal der Transformationsmittel entnommen und in den Grob-AFC-Mitteln mit einem festgelegten Daten-Schema korreliert werden, wobei die Art dieser Korrelation entsprechend dem aktuellen Modus gewählt wird;
- qualitative Auswertung der Ergebnisse der groben Zeitsynchronisation und der zur groben AFC-Korrektur gehörenden Korrelations-Ergebnisse, um die System-Konformität und Empfangsqualität der Digitalsignale zu bestimmen.

[0021] Vorteilhafte Weiterbildungen des erfindungsgemäßen Verfahrens ergeben sich aus den zugehörigen abhängigen Ansprüchen.

[0022] Im Prinzip ist der erfindungsgemäße Empfänger für Multicarrier-Digitalsignale, die in temporal-spektralen Rahmen angeordnet sind und Daten-Symbole mit einem Guardintervall und einer Nutzsymbollänge T_u und Referenzinformationen enthalten, und die in verschiedenartigen Modi übertragen werden können, versehen mit:

- Multipliziermitteln und Transformationsmitteln für das Digitalsignal;
- groben Zeit-Synchronisationsmitteln, in denen zur Abstimmung beim Empfang oder zur Überprüfung der Systemkonformität der empfangenen Signale das Digitalsignal in zeitlicher Richtung korreliert wird mit dem um verschiedene, den möglichen Modi entsprechende Werte von T_u zeitlich verschobenen Digitalsignal, wobei der aktuelle Modus aus der Lage und den Beträgen von Maxima der Korrelationswerte ermittelt wird und das aktuelle Guardintervall aus Abständen von Maxima der Korrelationswerte ermittelt wird und danach ein daraus sich ergebendes Abtastfenster mit einer T_u entsprechenden Länge für Transformationsmittel und die nachfolgende Signalauswertung gesetzt wird;
- Grob-AFC-Mitteln für vor den Transformationsmitteln angeordnete Multipliziermittel, in denen eine grobe AFC-Kor-

rektur mit Hilfe von dem Anordnungs-Schema der Referenzinformationen entsprechenden Informationen des aktuellen Symbols durchgeführt wird, die dem Ausgangssignal der Transformationsmittel entnommen und in den Grob-AFC-Mitteln mit einem festgelegten Daten-Schema korreliert werden, wobei die Art dieser Korrelation entsprechend dem aktuellen Modus gewählt wird;

- Auswerte-Mitteln zur qualitativen Auswertung der Ergebnisse der groben Zeit-Synchronisationsmittel und der in den Grob-AFC-Mitteln ermittelten Korrelations-Ergebnisse, die die System-Konformität und Empfangsqualität der Digitalsignale bestimmen.

[0023] Vorteilhafte Weiterbildungen des erfindungsgemäßen Empfängers ergeben sich aus den zugehörigen abhängigen Ansprüchen.

Zeichnung

[0024] Anhand der Zeichnung ist ein Ausführungs-Beispiel der Erfindung beschrieben.

Fig. 1 zeigt das Blockschaltbild für einen erfindungsgemäßen Empfänger.

Ausführungs-Beispiele

[0025] In dem Empfänger gemäß Fig. 1 wird zunächst für das digitale Eingangssignal INP in Grob-Zeitsynchronisations-Mitteln CTS eine grobe Synchronisation durchgeführt. Dabei wird das Zeitsignal mit dem um eine Nutzsymbollänge T_u verschobenen Zeitsignal, z.B. zweimal bis fünfmal pro Daten-Frame, korreliert. Bei dieser Korrelation werden Proben mit verschiedener Länge T_u entsprechend dem jeweiligen Mode verwendet und anhand der sich ergebenden, gefilterten bzw. gemittelten Korrelations-Ergebnis-Maxima wird dann in Mode-Detektormitteln MDET auf den aktuell vorliegenden Mode MO (z.B. 2K-oder 8K-Mode) geschlossen, z.B. durch Vergleich der Maxima mit einem gespeicherten Schwellwert. MDET gibt eine entsprechende Mode-Information MO ab.
Falls sich kein brauchbares Korrelations-Ergebnis-Maximum ergibt, können die Korrelationsschritte in CTS wiederholt werden. Aus dem Abstand der Korrelations-Maxima wird in CTS unter Berücksichtigung des Modes das verwendete Guardintervall ermittelt und nachfolgend ein Abtastfenster positioniert, z.B. durch einmaliges Setzen eines zur Symbolfolge (T_u+Δ) synchronen Zählers in CTS, der ein Zeitfenster mit Dauer T_u ausgibt, z.B. mittels eines Startsignals ST, welches Zeit-Feinsynchronisations-Mitteln FTS zugeführt wird. Mittels eines dabei benutzten Basis-Oszillators VCX0 werden in FTS die Position des Abtast-Fensters FFTWIN und der Abtast-Takt nachgeregelt.
Das aus einem I- und Q-Anteil bestehende Eingangssignal INP wird in einem Multiplizierer M mit einem aus einem Oszillator NCO stammenden Frequenz-Korrektursignal FCORR multipliziert. Das mit FFTWIN selektierte Ausgangssignal von M wird in Fast-Fourier-Transformationsmitteln FFT in den Frequenzbereich umgesetzt und bildet letztlich das aus einem I- und Q-Anteil bestehende Ausgangssignal OU.
Falls der Mode richtig erkannt und das Abtastfenster annähernd korrekt positioniert wurde, kann eine grobe AFC durch Grob-AFC-Mittel CAFC erfolgen. Dazu werden die vorgesehenen kontinuierlichen Pilotsignalen CPIL des aktuellen Symbols eines Datenrahmens dem Ausgangssignal von FFT entnommen und in CAFC mit einem festgelegten Schema korreliert (45 Positionen beim 2K-Mode, 177 Positionen beim 8K-Mode), und zwar über ±16 Shifts beim 2K-Mode bzw. ±64 Shifts beim 8K-Mode. Die Art der Korrelation wird entsprechend MO gewählt.
[0026] Zur Verbesserung der Grob-AFC können mehrere solcher Ergebnisse über eine bestimmte Anzahl, z.B. 3 bis 10, von Symbolen kombiniert oder gemeinsam verarbeitet werden, z.B. durch Mittelung, Majoritätsbildung oder Tiefpaß-Filterung. Das Maximum des Korrelationsvorgangs bzw. die aus mehreren solcher Maxima entsprechend abgeleitete Größe ergibt die grobe Frequenzabweichung $\Delta f = p'{}^*F_s$ und dient als Steuersignal für den Oszillator NCO. Die nächste Auswertung kann nach einem gewissen Abstand erfolgen, z.B. 3 bis 6 mal pro Rahmen. Wenn Δf einen vorgegebenen Wert D_{max} (z.B. $D_{max} = F_s/3$) unterschreitet, kann die entsprechende NCO-Abstimmung zunächst beibehalten werden und zur Fein-AFC in Fein-AFC-Mitteln FAFC übergegangen werden, denen ebenfalls die vorgesehenen kontinuierlichen Pilotsignalen CPIL des aktuellen Symbols zugeführt werden. Die Ausgangssignale von CAFC und FAFC werden in einem Komoinierer A kombiniert und als gemeinsames Steuersignal NCO zugeführt.
[0027] In einer Auswerte-Schaltung QREV werden die Korrelationsergebnisse aus CTS und CAFC qualitativ ausgewertet. Dazu erhält QREV ebenfalls die Mode-Information MO. Das Ausgangssignal RC von QREV steuert dann entsprechende Teile des Empfängers.
Nach Positionieren des Abtastfensters und/oder Erreichen von $\Delta f < D_{max}$ werden die obengenannten Bedingungen in bestimmten zeitlichen Abständen zum Zweck der Synchronisations-Kontrolle überprüft. Bei z.B. 2 bis 10-maligem negativem Ergebnis erfolgt ein Neustart mit der groben Zeit-Synchronisation in CTS.
[0028] Abhängig vom bisherigen Abstimmungs-Ergebnis wird das empfangene Signal im Empfänger als "systemkonform" bzw. "vorhanden" oder "nicht-systemkonform" bzw. "nicht vorhanden" gekennzeichnet. Je nach diesem aktuellen

Kennzeichnungs-Zustand wird beim Suchvorgang oder bei einer Empfangsprobe entweder die weitere Decodierung des Signals eingeleitet oder der Suchvorgang fortgesetzt oder bei der Empfangsprobe die Information "nicht vorhanden" ausgegeben.

Falls der Abstimmvorgang weitergeführt werden soll, kann nun eine Fein-AFC erfolgen. Dazu wird die Phasenänderung der kontinuierlichen Pilotsignalen CPIL von Symbol zu Symbol bestimmt und über sämtliche Pilotsignalen CPIL (45 beim 2K-Mode, 177 beim 8K-Mode) gemittelt. Diese Mittelwerte können tiefpaß-gefiltert werden und können, da sie proportional zu Δf sind, ebenfalls dem Oszillator NCO zugeführt werden, z.B. mittels Kombination in A, jedoch mit verminderter Steilheit.

[0029] Anschließend erfolgt eine Rahmen-Synchronisation und eine Zeit-Feinsynchronisation bzw. Abtast-Takt-Justierung. Dieses geschieht durch Auswertung der dem Ausgangssignal von FFT entnommenen TPS-Pilotzellen TPSPIL, die in einem TPS-Decoder TPSDEC decodiert werden. Dessen Ausgangssignal wird ebenfalls den Zeit-Feinsynchronisations-Mitteln FTS zugeführt und bewirkt eine entsprechende Nachregelung des Basis-Oszillators VCX0 zur Gewinnung des Abtast-Taktes CL sowie eine Korrektur der Position des Abtastfensters FFTWIN. Der Rahmenanfang (FTS-Ausgangssignal FTSO) und die Position der 'scattered Pilots' wird mit Hilfe der Sync-Sequenz der TPS-Pilotzellen durch Korrelation bestimmt. Der Abtast-Takt CL wird allen in Fig. 1 dargestellten Schaltungsteilen zugeführt.

[0030] Die 'scattered Pilots' können in FTS zeitlich so interpoliert werden, daß jeder dritte Träger als 'scattered Pilot' angesehen werden kann. Die Impuls-Antwort wird auf Basis der über die Zeit interpolierten 'scattered Pilots' mit Hilfe einer Division durch die spezifizierten 'scattered Pilots'Sollwerte und einer inversen FFT ermittelt.

Anschließend wird die Abweichung des Zentrums der ImpulsAntwort von einer für optimalen Empfang gewünschten Sollposition festgestellt. Dieser Vorgang wird vorteilhaft 3 bis 7 mal pro Rahmen wiederholt. Das Ergebnis wird vorteilhaft blockweise gefiltert und dann weiterverwertet. Aus der Größe und Richtung der so ermittelten Abweichung wird der Abtast-Takt-Referenzoszillator VCX0 in FTS nachgeregelt. Diese Nachregelung kann auch mittels Oszillator NCO und Multiplizierer M erfolgen. NCO kann eine digitale PLL enthalten.

[0031] Die Erfindung kann in DVB-Empfängern oder in Empfängern für vergleichbare digitale Signale zum Einsatz kommen, z.B. auch in DAB-Empfängern. Die angegebenen Zahlenwerte werden dann entsprechend geändert und die einzelnen Synchronisationsoder Überprüfungs-Schritte werden an die in den Rahmen aktuell übertragenen Referenz- oder Synchronisations-Daten angepaßt. Beim DAB-Empfänger wird dann anstelle des hier beschriebenen Grob-AFC-Korrelationsverfahrens (Basis sind kontinuierliche Pilotsignale) das in EP-A-0 786 889 beschriebene Verfahren (Basis sind CAZAC-Symbole) verwendet. Die qualitative Auswertung der erzielten Korrelationsergebnisse ist im Wesentlichen identisch. Die erfindungsgemäßen Auswertungen sind besonders vorteilhaft in kombinierten Empfängern (DAB und DVB-T oder digital und analog).

Patentansprüche

1. Verfahren für den Empfang von Multicarrier-Digitalsignalen (INP), die in temporal-spektralen Rahmen angeordnet sind und Daten-Symbole mit einem Guardintervall und einer Nutzsymbollänge T_u und Referenzinformationen (CPIL, SCPIL, TPSPIL) enthalten, und die in verschiedenartigen Modi (MO, 2k, 8k) übertragen werden können, **gekennzeichnet** durch folgende Schritte zur Abstimmung beim Empfang oder zur Überprüfung der Systemkonformität der empfangenen Signale:

   - grobe Zeitsynchronisation (CTS), bei der das Digitalsignal (INP) in zeitlicher Richtung korreliert wird mit dem um verschiedene, den möglichen Modi entsprechende Werte von T_u zeitlich verschobenen Digitalsignal (INP), wobei der aktuelle Modus (MO) aus der Lage und den Beträgen von Maxima der Korrelationswerte ermittelt wird und das aktuelle Guardintervall aus Abständen von Maxima der Korrelationswerte ermittelt wird und danach ein daraus sich ergebendes Abtastfenster mit einer T_u entsprechenden Länge für Transformationsmittel (FFT) und eine sich anschließende Signalauswertung gesetzt wird;
   - grobe AFC-Korrektur mit Hilfe von vor den Transformationsmitteln angeordneten Multipliziermitteln (M) und mit Hilfe von nach den Transformationsmitteln angeordneten Grob-AFC-Mitteln (CAFC), wobei dem Anordnungs-Schema der Referenzinformationen entsprechende Informationen (CPIL) des aktuellen Symbols dem Ausgangssignal der Transformationsmittel (FFT) entnommen und in den Grob-AFC-Mitteln mit einem festgelegten Daten-Schema korreliert werden, wobei die Art dieser Korrelation entsprechend dem aktuellen Modus (MO) gewählt wird;
   - qualitative Auswertung der Ergebnisse der groben Zeitsynchronisation (CTS) und der zur groben AFC-Korrektur gehörenden Korrelations-Ergebnisse, um die System-Konformität und Empfangsqualität der Digitalsignale (INP) zu bestimmem.

2. Verfahren nach Anspruch 1, wobei die Referenzinformationen kontinuierliche Pilotsignalen (CPIL), verstreute Pilotzellen (SCPIL) und TPS-Pilotzellen (TPSPIL) umfassen und bei der groben AFC-Korrektur kontinuierlichen Pilotsi-

gnalen (CPIL) entsprechende Informationen des aktuellen Symbols dem Ausgangssignal der Transformationsmittel (FFT) entnommen und in den Grob-AFC-Mitteln mit dem festgelegten Daten-Schema korreliert werden.

3. Verfahren nach Anspruch 1 oder 2, wobei, falls sich bei der groben Zeitsynchronisation (CTS) kein brauchbares Korrelations-Ergebnis-Maximum ergibt, die Korrelationsschritte wiederholt werden.

4. Verfahren nach einem oder mehreren der Ansprüche 1 bis 3, wobei jeweils mehrere Korrelations-Ergebnisse bei der groben AFC-Korrektur (CAFC) gemittelt oder gefiltert oder einer Majoritätsentscheidung zugeführt werden und wobei diese Ergebnisse auch bei der qualitativen Auswertung benutzt werden.

5. Verfahren nach einem oder mehreren der Ansprüche 1 bis 4, wobei bei der groben AFC-Korrektur (CAFC) aus jeweils mehreren zur Gruppe des Korrelations-Maximums gehörenden Werten die aktuelle Frequenzabweichung bestimmt wird.

6. Verfahren nach einem oder mehreren der Ansprüche 1 bis 5, wobei die Ergebnisse der qualitativen Auswertung zur Entscheidung darüber benutzt werden, ob die weitere Decodierung der Digital-Signale eingeleitet oder ein begonnener Signal-Suchvorgang mit neuer Frequenzeinstellung des Empfängers fortgesetzt werden soll.

7. Verfahren nach einem oder mehreren der Ansprüche 2 bis 6, wobei, falls die Frequenzabweichung eine festgelegte erste Schwelle unterschreitet, von der groben AFC-Korrektur (CAFC) auf eine Fein-AFC (FAFC) übergegangen wird, deren Steuergröße von der von Symbol zu Symbol festgestellten Phasenänderung der kontinuierlichen Pilotsignalen (CPIL) abgeleitet wird.

8. Verfahren nach einem oder mehreren der Ansprüche 2 bis 7, wobei beim laufendem Empfang, d.h. bei bereits erfolgter Decodierung der Digital-Signale, im Fall, daß die gesetzten Kriterien nicht mehr erfüllt sind, Maßnahmen wie ein Muting des Tons und/oder ein Einfrieren des Bildes in den nachfolgenden Empfängerstufen eingeleitet werden.

9. Verfahren nach einem oder mehreren der Ansprüche 2 bis 8, wobei nach der qualitativen Ergebnis-Auswertung eine Zeit-Feinsynchronisation durch Rahmen- und Abtasttakt-Synchronisationsmittel (FTS, TPSDEC) stattfindet, bei der eine Synchronisations-Sequenz aus dem Ausgangssignal der Transformationsmittel (FFT) entnommenen TPS-Pilotzellen (TPSPIL) dazu verwendet wird, die Rahmenposition und die Position der ebenfalls dem Ausgangssignal der Transformationsmittel (FFT) entnommenen verstreuten Pilotzellen (SCPIL) im Rahmen zu bestimmen und wobei die Abweichung des Zentrums einer mit Hilfe der verstreuten Pilotzellen gewonnenen Impuls-Antwort von einem Sollwert dazu verwendet wird, einen entsprechenden Abtast-Takt (CL) zu justieren.

10. Verfahren nach Anspruch 9, wobei bei der Zeit-Feinsynchronisation (FTS, TPSDEC) die verstreuten Pilotzellen (SCPIL) temporal interpoliert werden, um die Impuls-Antwort zu gewinnen.

11. Verfahren nach Anspruch 9 oder 10, wobei bei der Zeit-Feinsynchronisation (FTS, TPSDEC) die Abweichung vom Sollwert mehrfach festgestellt und diese Ergebnisse kombiniert werden.

12. Verfahren nach einem oder mehreren der Ansprüche 1 bis 11, wobei nach Positionieren des Abtastfensters und/oder Erreichen einer Frequenzabweichung, die kleiner ist als ein zweiter Schwellwert, die grobe AFC-Korrektur in bestimmten Abständen überprüft wird und, falls mehrfach die Frequenzabweichung größer ist als der zweite Schwellwert, ein Neustart der groben AFC-Korrektur erfolgt.

13. Empfänger für Multicarrier-Digitalsignale (INP), die in temporal-spektralen Rahmen angeordnet sind und Daten-Symbole mit einem Guardintervall und einer Nutzsymbollänge T_u und Referenzinformationen (CPIL, SCPIL, TPS-PIL) enthalten, und die in verschiedenartigen Modi (MO, 2k, 8k) übertragen werden können, versehen mit:

- Multipliziermittel (M) und Transformationsmit-tel (FFT) für das Digitalsignal (INP);
- grobe Zeit-Synchronisationsmittel (CTS), in denen zur Abstimmung beim Empfang oder zur Überprüfung der Systemkonformität der empfangenen Signale das Digitalsignal (INP) in zeitlicher Richtung korreliert wird mit dem um verschiedene, den möglichen Modi entsprechende Werte von T_u zeitlich verschobenen Digitalsignal (INP), wobei der aktuelle Modus (MO) aus der Lage und den Beträgen von Maxima der Korrelationswerte ermittelt wird und das aktuelle Guardintervall aus Abständen von Maxima der Korrelationswerte ermittelt wird

und danach ein daraus sich ergebendes Abtastfenster mit einer T_u entsprechenden Länge für Transformationsmittel (FFT) und die nachfolgende Signalauswertung gesetzt wird;

- Grob-AFC-Mitteln (CAFC) für vor den Transformationsmitteln angeordnete Multipliziermittel (M), in denen eine grobe AFC-Korrektur mit Hilfe von dem Anordnungs-Schema der Referenzinformationen entsprechende Informationen (CPIL) des aktuellen Symbols durchgeführt wird, die dem Ausgangssignal der Transformationsmittel (FFT) entnommen und in den Grob-AFC-Mitteln mit einem festgelegten Daten-Schema korreliert werden, wobei die Art dieser Korrelation entsprechend dem aktuellen Modus (MO) gewählt wird;
- Auswerte-Mittel zur qualitativen Auswertung der Ergebnisse der groben Zeit-Synchronisationsmittel (CTS) und der in den Grob-AFC-Mitteln (CAFC) ermittelten Korrelations-Ergebnisse, die die System-Konformität und Empfangsqualität der Digitalsignale (INP) bestimmem.

14. Empfänger nach Anspruch 13, wobei die Referenzinformationen kontinuierliche Pilotsignalen (CPIL), verstreute Pilotzellen (SCPIL) und TPS-Pilotzellen (TPSPIL) umfassen und für die Grob-AFC-Mittel (CAFC) kontinuierliche Pilotsignalen (CPIL) des aktuellen Symbols dem Ausgangssignal der Transformationsmittel (FFT) entnommen und in den Grob-AFC-Mitteln mit dem festgelegten Daten-Schema korreliert werden.

15. Empfänger nach Anspruch 13 oder 14 mit Rahmen- und Abtasttakt-Synchronisationsmitteln (FTS, TPSDEC), in denen eine Zeit-Feinsynchronisation durchgeführt wird durch Auswertung von einer Synchronisations-Sequenz aus dem Ausgangssignal der Transformationsmittel (FFT) entnommenen TPS-Pilotzellen (TPSPIL), um die Rahmenposition und die Position der ebenfalls dem Ausgangssignal der Transformationsmittel (FFT) entnommenen verstreuten Pilotzellen (SCPIL) im Rahmen zu bestimmen, wobei die Abweichung des Zentrums einer mit Hilfe der verstreuten Pilotzellen gewonnenen Impuls-Antwort von einem Sollwert ausgewertet wird, um einen Abtast-Takt (CL) zu justieren.

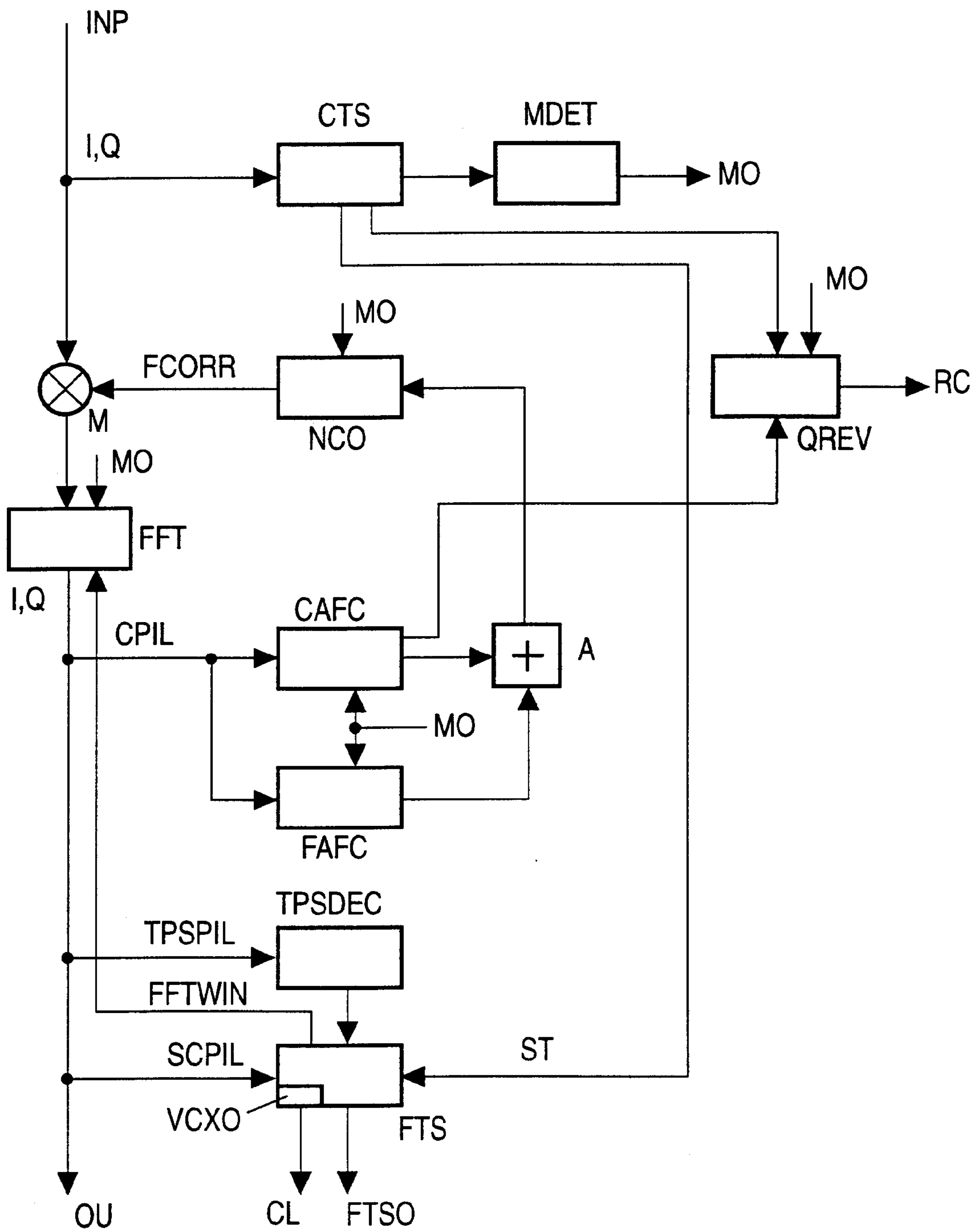
INP
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Fig.1

Europäisches Patentamt

EUROPÄISCHER RECHERCHENBERICHT

Nummer der Anmeldung
EP 97 11 2929

EINSCHLÄGIGE DOKUMENTE

Kategorie	Kennzeichnung des Dokuments mit Angabe, soweit erforderlich, der maßgeblichen Teile	Betrifft Anspruch	KLASSIFIKATION DER ANMELDUNG (Int.Cl.6)
A	WO 97 26742 A (FRANCE TELECOM; TELEDIFFUSION DE FRANCE; HELARD ET AL.) * Zusammenfassung * * Seite 2, Zeile 5 - Zeile 13 * * Seite 3, Zeile 3 - Zeile 13 * * Seite 4, Zeile 27 - Zeile 29 * ---	1-15	H04L27/26 H04L27/00 H04H1/00
A	DE 44 03 408 C (GRUNDIG) * Zusammenfassung * * Spalte 1, Zeile 48 - Zeile 58 * ---	1,13	
A	EP 0 562 422 A (GENERAL INSTRUMENT CORPORATION) * Zusammenfassung * ---	1,13	
A	WO 97 07620 A (PHILIPS; PHILIPS) * Zusammenfassung; Abbildung 3 * * Seite 2, Zeile 26 - Zeile 39 * ---	1-15	
A	EP 0 653 858 A (TOSHIBA) * Zusammenfassung * * Spalte 2, Zeile 39 - Zeile 52 * * Spalte 3, Zeile 27 - Zeile 31 * ---	1-15	RECHERCHIERTE SACHGEBIETE (Int.Cl.6) H04L
E	EP 0 786 889 A (THOMSON BRANDT) * Seite 3, Zeile 20 - Zeile 37 * * Seite 3, Zeile 57 - Zeile 59 * * Seite 4, Zeile 19 - Zeile 25 * -----	1-15	

Der vorliegende Recherchenbericht wurde für alle Patentansprüche erstellt

Recherchenort	Abschlußdatum der Recherche	Prüfer
DEN HAAG	9.Dezember 1997	Scriven, P

KATEGORIE DER GENANNTEN DOKUMENTE

X : von besonderer Bedeutung allein betrachtet
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(71) Applicant:
LUCENT TECHNOLOGIES INC.
Murray Hill, New Jersey 07974-0636 (US)

(72) Inventor: Van Nee, C.J. Richard
3454 CG de Meern (NL)

(74) Representative:
Watts, Christopher Malcolm Kelway, Dr.
Lucent Technologies (UK) Ltd,
5 Mornington Road
Woodford Green Essex, IG8 0TU (GB)

(54) Multicarrier modulation system, with variable symbol rates

(57) An OFDM system uses a normal mode which has a symbol length T, a guard time T_G and a set of N sub-carriers, which are orthogonal over the time T, and one or more fallback modes which have symbol lengths KT and guard times KT_G where K is an integer greater than unity. The same set of N sub-carriers is used for the fallback modes as for the normal mode. Since the same set of sub-carriers is used, the overall bandwidth is substantially constant, so alias filtering does not need to be adaptive. The Fourier transform operations are the same as for the normal mode. Thus fallback modes are provided with little hardware cost. In the fallback modes the increased guard time provides better delay spread tolerance and the increased symbol length provides improved signal to noise performance, and thus increased range, at the cost of reduced data rate.

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Description

Technical Field

[0001] This invention relates to communication systems and, more particularly, OFDM (Orthogonal Frequency Division Multiplexing) modulation schemes.

Background of the Invention

[0002] OFDM is a block-oriented modulation scheme that maps N data symbols into N orthogonal sub-carriers separated by a frequency interval of 1/T, where T is the symbol duration, i.e. the time period over which the sub-carriers are orthogonal. As such, multi-carrier transmission systems use OFDM modulation to send data bits in parallel over multiple sub-carriers (also called tones or bins). An important advantage of multi-carrier transmission is that inter-symbol interference due to signal dispersion (or delay spread) in the transmission channel can be reduced or even eliminated by inserting a guard time interval T_G between the transmission of subsequent symbols, thus avoiding an equaliser as required in single carrier systems. This gives OFDM an important advantage over single carrier modulation schemes. The guard time allows delayed copies of each symbol, arriving at the receiver after the intended signal, to die out before the succeeding symbol is received. OFDM's attractiveness stems from its ability to overcome the adverse effects of multi-channel transmission without the need for equalisation.

[0003] The transformations between blocks of symbols and base-band carrier signal are normally carried out using fast Fourier transform (FFT) techniques. A discussion of OFDM is given by Alard and Lasalle, EBU Technical Review, no. 224, August 1987, pages 168-190.

[0004] A need exists for a flexible OFDM system which provides the advantages of OFDM to a variety of communication environments.

[0005] In a previous patent application (US, serial No. 08/834684, herein referred to as VN) I disclosed several techniques to scale data rates using OFDM. Scaling methods involve changing the clock rate, FFT size, coding rate, constellation size and guard time.

[0006] The present invention is intended to provide fallback rates with a minimum change in hardware.

Summary of the Invention

[0007] The invention is as set out in the independent claims, preferred forms being set out in the dependent claims.

[0008] In a preferred embodiment of the present invention, a first signalling mode (the 'normal' mode) uses a symbol length T, a guard time T_G and a set of N sub-carriers and a second mode (the 'fallback' mode) uses a symbol length KT, a guard time KT_G and the same set of N sub-carriers, where K is an integer greater than unity.

[0009] The technique can increase the range and delay spread tolerance without substantially changing the bandwidth and without changing the FFT size, at the cost of a decreased bit rate. Further, the fallback rates can also be used to provide a multiple access capability, so using fallback rates does not necessarily result in a bad spectral efficiency.

Brief Description of the Drawings

[0010]

Figures 1 and 2 illustrate the transmission of an OFDM symbol in K = 1 mode and K = 2 mode according to the invention,

Figure 3 shows, in block schematic form, a transmitter embodying the invention; and

Figure 4 shows, in block schematic form, a receiver embodying the invention.

[0011] Figure 1 shows an OFDM symbol transmitted with a symbol duration T and a guard time T_G. The object of the guard time T_G is to accommodate any interference between consecutive symbols due to dispersion or multi-path interference (collectively referred to as 'delay spread'), and to leave a time T over which the symbol can be received free from such interference. Under some conditions, or in some applications, it may happen that the guard time T_G is insufficient to accommodate this delay spread (as in Figure 1). It may also happen that a greater range will be required, i.e. a higher signal-to-noise ratio in the recovered signal. Simply increasing the guard time T_G would accommodate a larger delay spread, though it would not affect the range. Decreasing the clock rate seems a simple way of increasing the guard time T_G and the symbol duration T, but it would also decrease the frequency spacing 1/T between the sub-carriers. This would proportionately decrease the overall bandwidth of the channel, which would mean that the filters that are required to remove aliases would have to be adaptable, thus increasing the hardware requirement.

[0012] Figure 2 shows a symbol which has been transmitted with twice the symbol duration 2T and with twice the guard time $2T_G$. The guard time is now doubled, and can accommodate the illustrated intersymbol interference. Also, since the symbol duration is doubled, the signal-to-noise performance, and hence the range, is improved. It is important to note that the frequencies of the sub-carriers are not also halved as would be the case with a simple halving of the clock rate. The same set of sub-carriers is used, still separated by 1/T, not 1/2T. Therefore, the overall bandwidth of the channel, which is mainly determined by the spread of sub-carrier frequencies, and only to a much lesser extent by the widths of the individual sub-carriers, is substantially

unchanged.

[0013] Since for any OFDM symbol, the signal repeats itself after T seconds, where T is the FFT interval, it is possible to do 2 FFTs on two different parts of the received symbol, each with a length of T seconds. Since both FFT outputs carry the same data, but different noise, they can be averaged to get a 3 dB increase in signal-to-noise ratio. The FFT is a linear operation, so it is also possible to first average two T seconds intervals and use this averaged signal as input to a single FFT. This scheme can easily be extended to other data rates; in general, any rate which is a factor K less than the highest bit rate can be produced by extending the symbol duration by a factor of K. By taking K FFTs per symbol, a processing gain of K is achieved which increases the range. At the same time, the delay spread tolerance is increased by a factor of K. The only extra hardware required is for averaging K consecutive signal intervals of T seconds. In fact, the amount of processing in terms of operations per second is decreased for fallback rates, because the averaging takes far less processing than the FFT. Consider, for instance, the case of an OFDM modem with a 64 point FFT and a symbol duration of 2µs. A 64 point FFT involves about 192 complex multiplications and additions, so the procesing load is 96 Mops, where an operation is defined as one complex multiply plus one addition. If the symbol duration is doubled to create a fallback rate, then in 4µs, 64 additions have to be performed plus one 64 point FFT. Thus, the processing load becomes (192+64)/4µs = 64 Mops. In fact, this figure is pessimistic, because the extra additions have been given the same weight as multiplications, while they are significantly less complex when implemented in hardware. The additions are the only part of the receiver that has to run at the full clock rate; the FFT and everything following the FFT (channel estimation, decoding) can run at a rate that is K times lower than the original rate, which helps to reduce the power consumption.

[0014] Figure 3 shows an OFDM transmitter which receives a stream of data bits. A coding circuit 1 receives the data stream and partitions it into successive groups or blocks of bits. The coding circuit 1 introduces redundancy for forward error correction coding.

[0015] The blocks of coded data bits are input into a N-points complex IFFT (Inverse Fast Fourier Transform) circuit 2 where N is the number of the OFDM sub-carriers. In this particular embodiment, using quaternary phase-shift keying (QPSK), the IFFT is performed on blocks of 2N coded data bits received from the coding circuit 1. In practice, the transmitter has to use oversampling to produce an output spectrum without aliasing which introduces unwanted frequency distortion due to (intended or unintentional) low pass filtering in subsequent stages of the transmitter or in the transmission channel. Thus, instead of a N-points IFFT an M-points IFFT is actually done where M>N to perform the oversampling. These 2N bits are converted into N complex numbers, and the remaining M-N input values are set to zero.

[0016] To decrease the sensitivity to inter-symbol interference, the cyclic prefixer and windowing block 3 copies the last part of the OFDM symbol and augments the OFDM symbol by prefixing it with the copied portion of the OFDM symbol. This is called cyclic prefixing. Control circuitry 4 controls the cyclic prefixer and windowing block 3 to switch the guard time and the symbol duration as required, or as appropriate, between their normal values T_G and T respectively and their fallback values KT_G and KT respectively. To provide the fallback values the cyclic prefixer has to augment the OFDM symbol with K-1 copies of itself, in addition to the prefix, which is preferably K times as long as the normal prefix.

[0017] To reduce spectral sidelobes, the cyclic prefixing and windowing block 3 performs windowing on the OFDM symbol by applying a gradual roll-off pattern to the amplitude of the OFDM symbol. The OFDM symbol is input into a digital-to-analogue converter after which it is sent to a transmitter front-end 6 that converts the baseband wave form to the appropriate RF carrier frequency in this particular embodiment for transmission from antenna 7.

[0018] With particular reference to Figure 4, the transmitted OFDM signal is received by an OFDM receiver through an antenna 10. The OFDM signal is processed (down-converted) using the receive circuitry 11. The processed OFDM signal is input into a analog-to-digital converter 12. The digital OFDM signal is received by a symbol timing circuit 13 which acquires the OFDM symbol timing and provides a timing signal to a Fast Fourier Transform (FFT) block 14 and an integrate and dump filter 15. The integrate and dump filter 15 sums K samples that are separated by T seconds. The memory of the filter — which consists of a delay line of M samples, where M is the FFT size — is cleared at the start of each new symbol. This reset time is indicated by the timing circuit 13 which is already present in a normal OFDM receiver to indicate the start of the FFT interval. A control circuit 16 sets the number of averaging intervals K.

[0019] As an alternative implementation, the integrate and dump filter could be placed after the FFT circuit 14 instead of before. In that case, for each symbol, K consecutive FFT outputs are averaged. However, the processing load is increased because the FFT always has to run at the maximum clock rate.

[0020] The sequence of symbols produced by the FFT circuit 14 is applied to conventional decoding circuitry 17 to produce the data output signal.

[0021] When a fallback rate is used at a rate that is K times lower than the original rate, the above described technique will produce subcarriers each of which has a bandwidth that is K times smaller than the original bandwidth. Thus, although the total signal bandwidth does not substantially change, the bandwidth of each subcarrier does become smaller. This makes it possible to do frequency division multiple access of up to K users in

the same band. Each user has to shift its carrier frequency by a different multiple of 1/KT in order to stay orthogonal to the other users. As a example, when 64 subcarriers are used with a subcarrier spacing of 1 MHz, then it is possible to accommodate 4 users in the same channel when using a fallback rate with K=4. All 4 users use the same transmission and reception scheme as described above, but their carrier frequencies have an offset of 0, 250, 500 and 750 KHz, respectively, or, in general, n/KT, where the values of n are different MODULO K.

[0022] As discussed in VN, the control circuits 4, 16 may be responsive to external settings and/or the results of monitoring the signal quality. As also discussed in VN, it may be appropriate to use different modes for the up-links and the down-links in a communications system.

Claims

1. Orthogonal frequency division multiplex communications apparatus employing a set of sub-carriers which are orthogonal over a time T, information-carrying symbols being expressed by superpositions of said sub-carriers,
CHARACTERISED IN THAT
the apparatus is configured to selectively operate in one of a plurality of signalling modes in each of which the duration of each said symbol is KT where K is a positive integer, different ones of the said modes having different values of K but the same set of sub-carriers.

2. Apparatus as claimed in claim 1 wherein one of the said modes has K=1.

3. Apparatus as claimed in claim 1 or claim 2 wherein a guard time is interposed between successive ones of said symbols, the length of said guard time being greater for modes with a greater value of K.

4. Apparatus as claimed in claim 3 wherein the length of said guard time is KT_G where T_G is the same for all of the said modes.

5. Apparatus as claimed in any of the preceding claims, being a receiver and including Fourier transform means (14) for recovering said symbols from said superposition of sub-carriers and averaging means (15) for providing, when operating in a mode in which K>1, an average over K successive periods of duration T.

6. Apparatus as claimed in claim 5 wherein said averaging means (15) are connected upstream of the Fourier transform means (14) to receive the superposition of subcarriers of duration KT and derive an averaged superposition as input to the Fourier transform means (14).

7. Apparatus as claimed in any of claims 1 to 4, being a transmitter and including means (3,4) arranged to receive the superpositions of sub-carriers expressing the symbols and to derive a K-fold repetition of each said superposition.

8. A method of signalling using orthogonal frequency division multiplexing employing a set of sub-carriers which are orthogonal over a time T, information-carrying symbols being expressed by superpositions of said sub-carriers,
CHARACTERISED BY
selecting one of a predetermined plurality of signalling modes in each of which the duration of each said symbol is KT where K is a positive integer, different ones of the said modes having different values of K, but the same set of sub-carriers.

9. A method as claimed in claim 8 wherein one of the said modes has K=1.

10. A method as claimed in claim 8 or claim 9 wherein a guard time is interposed between successive ones of said symbols, the length of said guard time being greater for modes with a greater value of K.

11. A method as claimed in claim 10 wherein the length of said guard time is KT_G where T_G is the same for all of the said modes.

Fig.1.

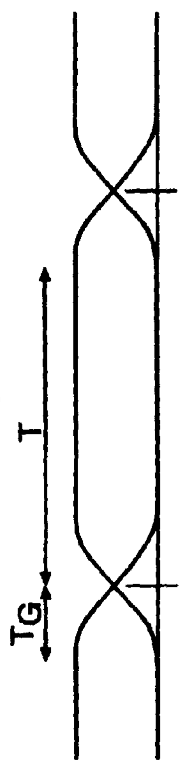


Fig.2.

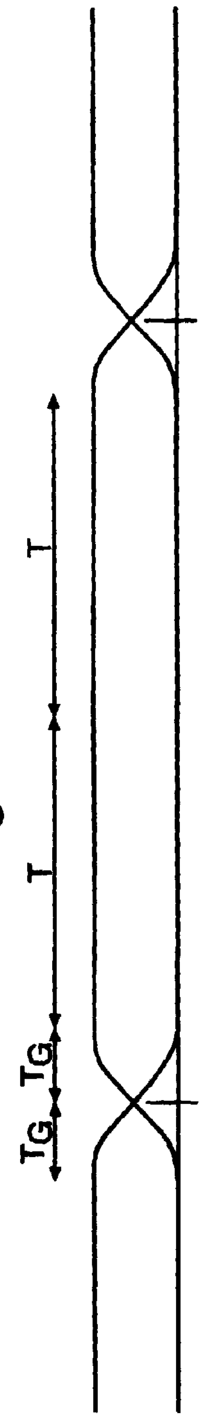


Fig.3.

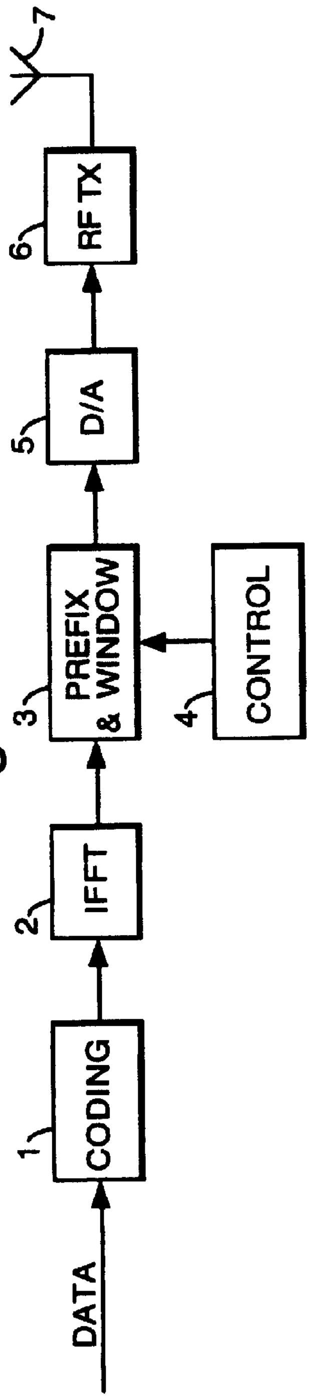
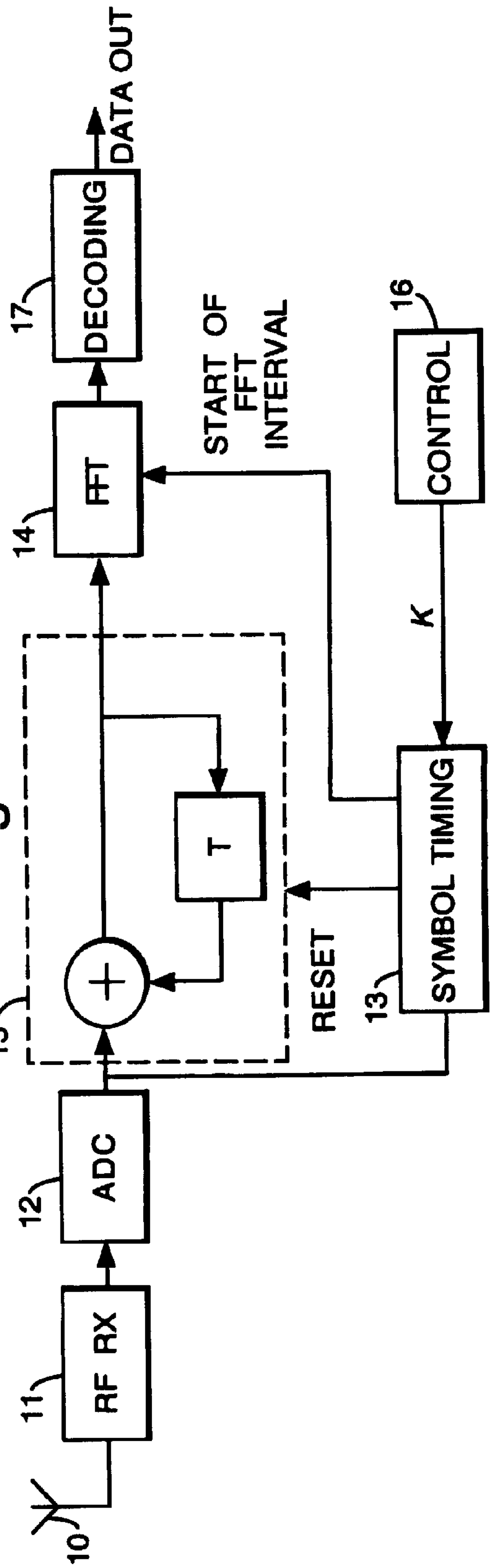

DATA
CODING
1
IFFT
2
PREFIX & WINDOW
3
CONTROL
4
D/A
5
RF TX
6
7

Fig.4.


10
RF RX
11
ADC
12
15
T
RESET
SYMBOL TIMING
13
K
CONTROL
16
FFT
14
START OF FFT INTERVAL
DECODING
17
DATA OUT

European Patent Office

EUROPEAN SEARCH REPORT

Application Number
EP 98 20 0010.1

DOCUMENTS CONSIDERED TO BE RELEVANT

Category	Citation of document with indication, where appropriate, of relevant passages	Relevant to claim	CLASSIFICATION OF THE APPLICATION (Int.Cl.6)
A	LARSSON ET AL.: "Mixed traffic in a multi carrier system" 1996 IEEE VEHICULAR TECHNOLOGY CONFERENCE, 28 April 1996 - 1 May 1996, NEW YORK, US, pages 1259-1263, XP000593147 * page 1262, right-hand column, paragraph 3 - paragraph 6 *	1,8	H04L27/26 H04L1/12
A	EP 0 589 709 A (MATSUSHITA) 30 March 1994 * page 45, line 20 - line 38 *	1,8	
A	RASMUSSEN ET A.: "A unifying discrete-time model for direct sequence and multicarrier variable rate broadband CDMA" SEVENTH IEEE INTERNATIONAL SYMPOSIUM ON PERSONAL, INDOOR AND MOBILE RADIO COMMUNICATIONS., 15 - 18 October 1996, NEW YORK, US, pages 1111-1115 vol.3, XP002068805 * page 1112, left-hand column *	1,8	
			TECHNICAL FIELDS SEARCHED (Int.Cl.6)
A	OKADA ET AL.: "Wideband indoor radio system using orthogonal multicarrier modulation" IEEE INTERNATIONAL CONFERENCE ON SYSTEMS ENGINEERING, 17 - 19 September 1992, NEW YORK, US, pages 457-462, XP002068806 * page 457, right-hand column, paragraph 4 *	1,8	H04L

The present search report has been drawn up for all claims

Place of search	Date of completion of the search	Examiner
THE HAGUE	19 June 1998	Scriven, P

CATEGORY OF CITED DOCUMENTS

X : particularly relevant if taken alone
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(71) Applicant: **LUCENT TECHNOLOGIES INC.**
Murray Hill, New Jersey 07974-0636 (US)

(72) Inventors:
- **Foschini, Gerard Joseph**
  South Amboy, NJ 08879 (US)
- **Golden, Glenn David**
  Boulder, Colorado 80302 (US)

(74) Representative:
Johnston, Kenneth Graham et al
Lucent Technologies (UK) Ltd,
5 Mornington Road
Woodford Green Essex, IG8 OTU (GB)

(54) Wireless communications system having a space-time architecture employing multi-element antennas at both the transmitter and receiver

(57) The bit rate at which a digital wireless communications system communicates data may be significantly increased by using multiple antennas at both the transmitter and receiver and by decomposing the channel into m subchannels that operate in the same frequency band. The system transmits m one dimensional signals and maximizes the minimum signal-to-noise ratio of the receiver detection process.

EP 0 951 091 A2

Description

Technical Field

[0001] The invention relates to a wireless communications system employing Multi-Element Antennas (MEAs) at both the transmitter and receiver.

Background of the Invention

[0002] The transmission capacity (ultimate bit rate) of a digital wireless communications system is based on a number of different parameters including (a) total radiated power at the transmitter, (b) the number of antenna elements at the transmitter and receiver, bandwidth, (c) noise power at the receiver, (d) characteristics of the propagation environment, etc. For a wireless transmission system employing an appreciable number of antennas at both the transmitter and receiver and operating in a so-called Rayleigh fading environment even without coding, the bit rate could be very large, e.g., 36 bits per second per Hz with a reasonable Signal to Noise Ratio (SNR) of 18 dB. Heretofore, it was difficult for a communications system to exchange data at a fraction of such a rate. The main reason for this is that the prior art did not appreciate the problems that had to be solved in order to build a large bit rate system.

Summary of the Invention

[0003] We have recognized that a wireless communications system that transmits at a substantial bit rate may be achieved, in accordance with an aspect of the invention, by decomposing an m-dimensional system into m-one dimensional systems (of possibly a different capacity) when the transfer (H matrix) characteristics of the wireless channel are unknown to the transmitter. More specifically and in accordance with the principles of the invention, a burst of signal vectors is formed from different data symbols and then transmitted by a transmitter via a Multi-Element Antenna Array. The transmitted vector symbols are received as signal vectors by a plurality of different antennas associated with a wireless receiver. The symbol components of the transmitted vector symbol have an (arbitrary) order and the receiver determines the best reordering of these transmitted components and then processes the received vector to determine the reordered transmitted symbol components. Such processing starts with the lowest (e.g., first) level of the reordered components, and for each such level, cancels out interfering contributions from lower levels, if any, and nulls out interfering contributions from higher levels, if any.

[0004] Such receiver processing includes, in accordance with one embodiment of the invention, compensating the weaker of the received transmitted signal components by first removing the interference from the stronger of the received transmitted components, and then processing the result to form the bit decisions.

[0005] These and other aspects of the invention may be appreciated from the following detailed description, accompanying drawings and claims.

Brief Description of the Drawings

[0006] In the drawings:

FIG. 1 illustrates in block diagram form a wireless transmitter and receiver embodying the principles of the invention;
FIG. 2 illustrates a more detailed block diagram of the wireless receiver of FIG. 1;
FIG. 3 illustrates the reordered transmit components and corresponding decision statistics for one of the n-dimensional complex signal vectors of a received burst of κ vectors;
FIG. 4. illustrates graphically the way in which such processing is achieved to eliminate interfering signals from a signal vector being processed by the receiver of FIG. 2;
FIGs. 5 and 6 illustrate in flow chart form the program which implements such reordering in the processor 60 of FIG. 2; and
FIGs. 7 and 8 illustrate in flow chart form the processor program which processes each received vector signal to determine the corresponding transmit symbols.

Detailed Description

[0007] The following illustrative embodiment of the invention is described in the context of a point-to-point communication architecture employing a transmitter having an array of m antenna elements and a receiver having an array of n antenna elements, where, in an illustrative embodiment of the invention, $m \le n$ and both values are > 1. For example, for transmissions at 1.9 Hz, $m \le 16$ and $n = 16$. Moreover, as will be seen below, the bit rate/bandwidth, in

terms of the number of bits/cycle, that may be achieved using the inventive architecture is indeed large.

[0008] For the sake of clarity, the detailed description assumes the following conditions. Specifically, assume that the transmitter and receiver, FIG. 1, are spaced very far apart, for example, 100λ. Also, the volume of such transmit and receive spaces is sufficient to accommodate one or more signal decorrelations. Also, locations of the transmitter and receiver may be reversed and several antennas would still be available at the receiver to receive substantially spatially decorrelated electromagnetic field samples originating from electromagnetic waves launched by the transmitter. This randomly faded matrix channel characteristic is assumed not to be known to the transmitter but may be "learned" by the receiver using certain channel measurements. Such training may be done by, for example, using m applications of standard n-fold receive diversity, one such application per transmit antenna. One example of this procedure is disclosed in the article entitled "Signal Acquisition and Tracking with Adaptive Arrays in the Digital Mobile Radio System IS-54 with Flat Fading, by J. H. Winters and published in the IEEE Transactions on Vehicular Technology, November 1993, which is hereby incorporated by reference.

[0009] The following description is also discussed in the context of burst mode communications, in which temporal changes in the channel are assumed to be negligible over the duration of a burst of data, and in which the characteristics (i.e., transmission environment) of the channel may change from burst to burst. Because of this, the channel bit rate that may be achieved is treated as a random variable. Along with (m, n), a key system parameter is the spatial average Signal-to-Noise Ratio (SNR), p, as would be measured using a "probe" antenna element at the transmitter end and at the receiver end of the transmission volumes. The total radiated power is considered to be constrained so that if m, for example, is increased, there is proportionately less power per transmit antenna. (Note, that in a Rayleigh propagation environment, p is independent of the number of transmit antennas).

[0010] As mentioned above, the transmission environment of a channel is assumed to be random, and time is assumed to be discrete. Also assume the following notation:

[0011] The transmitted signal is designated as s(t) and the total power is constrained to P_o (defined below) regardless of the value of m (the dimension of s(t)). For the sake of simplicity, the following discussion assumes that the bandwidth is sufficiently narrow so that the response of the channel may be considered to be flat over the channel frequency band.

[0012] The noise signal at the receiver is designated as ν(t), and is a complex n-D (Dimensional) Additive White Gaussian Noise (AWGN) process with statistically independent components of identical power N (one component for each of the n receiver antennas).

[0013] The received signal is designated as r(t), and, at a particular point in time, a received n-D signal includes one complex component per receive antenna. In the case where there is only one transmit antenna, then the transmitter radiates power P_o and the resulting (spatial) average power at the output of any of the receiving antennas is noted by P.

[0014] The spatial average SNR, ρ, at each receiver antenna is considered to be equal to P/N and independent of m in the Rayleigh propagation case.

[0015] The duration of a burst of data is κ vector symbols and is assumed to be equal to the number of intervals ("ticks") of a discrete time clock that occurs in such a burst.

[0016] A so-called matrix channel impulse response, g(t), has m columns and n rows. G(f) is used for the Fourier transform of g(t). To be consistent with the narrowband assumption, this matrix/transform is considered to be constant over the band of interest, in which notation G indicates that frequency dependence is suppressed. Thus, except for g (0), g(t) is the zero matrix. As will be seen, it may be convenient to represent the response of the matrix channel in a normalized form, h(t). Also, related to G, and for a matrix H, the equation $P^{1/2}.G = P_o^{1/2}.H$ defines the relationship between G and H, which provide $g(t) = (P_o/P)^{1/2}.h(t)$.

[0017] Further, realizations of a form of H for an ideal Rayleigh propagation environment may be modeled so that the m x n matrix entries are the result of independent identically distributed complex Gaussian variables of unit variance.

[0018] With the foregoing in mind, and using * to designate convolution, the basic vector equation describing the channel environment affecting a transmitted signal may be represented as follows:

$$r(t)= g(t) * s(t) + \nu(t) \qquad (1)$$

[0019] The two vectors added on the right hand side of equation (1) are complex n-D(dimensional) vectors (i.e., 2n real dimensions). For the above narrowband assumption, equation 1 may be simplified by replacing the convolution using a matrix-vector product as follows:

$$r(t)= (P_o/(P \cdot m))^{1/2} \cdot h(t) \cdot s(t) + \nu(t) \qquad (2)$$

[0020] A broad block diagram illustrating a generalized version of a communication system that processes the received vector signal described by equation 2 is shown FIG. 1. In particular, source 50 supplies a m-dimensional signal

to transmitter processor 100 which then transmits over selected ones of the antennas 110-1 through 110-k a m-dimensional symbol generated as a result of using m instances of a particular modulation technique, e.g., Quadrature Amplitude Modulation (QAM), in which each symbol component corresponds to a group of sequential bits. In an illustrative embodiment of the invention, the selection of m transmit antennas may be arbitrary. However, the selection could turn out to be inferior. Rather than being "stuck" with such a selection, transmitter processor 100 is arranged to systematically (or randomly) change the selection of antennas to avoid being "stuck" with an inferior selection for any appreciable amount of time. (It is noted that in one embodiment of the instant invention, a so-called feedback channel from receiver 200 to transmitter 100 is provided. The selection of transmitter antennas is then somewhat optimized based on information provided by receiver 200 to transmitter 100 via the feedback channel (represented in FIG. 1 by the dashed line.) Transmitter processor 100, more particularly, may be arranged to initially match the sequential order in which symbols are generated with the order of antennas 110-1 through 110-k, in which the first of such symbols is transmitted over antenna 110-1, the second symbol is then transmitted over antenna 110-2, and so on. If that selection turns out to be inferior based on the receiver feedback information, then transmitter processor 100 may change the selection or use a subset of the transmit antennas 110-I, all in accordance with an aspect of the invention. For example, as a result of such feedback, if the transmitter "learns" that the environment of the channel over which antennas 110-k-1 and 110-k transmit, then processor 100 may use just a subset of the antennas, e.g., 110-1 through 110-k-2, and select those antenna that may possibly result in the best reception at receiver 200 as reported via the feedback channel.)

[0021] For the system of FIG. 1, m different QAM signals may be considered to be statistically independent, but are otherwise identical modulations (although different modulations are allowable). (Note that each of the QAM modulated components of a received vector will also be referred to herein as a substream.) Also for expositional convenience, q (t) may be defined as follows:

$$q(t) \triangleq (P_o/(P \cdot m))^{1/2} \cdot s(t) \qquad (3)$$

[0022] Using equation (3), equation (2) may then be expressed simply as

$$r(t) = h(t) \cdot q(t) + \nu(t) \qquad (4)$$

The received vector, r(t), is used to estimate the m QAM components of the vector q(t), in which n components of the signal vector, r(t), are received by each of the receiver 200 antennas 120-1 through 120-n, respectively. The processing of the received signal to detect the transmitted symbols is the problem that receiver 200 needs to solve in accordance with the principles of the invention. The detected symbols are then mapped to respective bit sequences to reconstitute the original bit stream.

[0023] Note, that for the sake of clarity and convenience, the following discussion relating to the processing of a received vector suppresses the argument (t). Thus, for example, r(t) will be referred to simply as r, q(t) as q, etc.

[0024] Receiver 200, FIG. 2, more particularly includes, inter alia, a bank of conventional RF receiver sections (not shown) which respectively interface with antennas 120-1 through 120-n. It also includes preprocessor (also just processor) 60, symbol processor 65 and multiplexer 70. Preprocessor 60 receives the signals as respective signal vectors from the n antennas 120-1 through 210-n, and preprocesses each received signal vector to eliminate interference between the signal components forming that vector. Such processing includes (a) subtracting interference stemming from previously detected transmitted symbols from the vector being processed, (b) nulling out of the vector being processed interference from other transmitted symbols that have not yet been processed and detected, and (c) using the stronger elements of the received signal to compensate for the weaker elements, all in accordance with the principles of the invention and as will be discussed below in detail. (Such processes are illustrated in FIG. 4 as procedures 40-1 through 407.) In accordance with an aspect of the invention, such compensation is achieved by determining the best reordering of the transmitted components for their detection and then processing the received vector to determine the reordered transmitted symbol components.

[0025] The reordering of the transmit components may be achieved by, for example, placing at the bottom of the vertical stack (level (1)) the detection process which will estimate the transmitted signal component that offers the largest SNR, then placing next in the vertical stack (level (2)) the detection process that will estimate the transmitted signal component having the next largest SNR of the remaining m-1 transmitted components and so on, as will be explained below in detail. (It can be appreciated from the foregoing that an alternative stacking arrangement could be readily used without departing from the spirit and scope of the instant invention)

[0026] A received vector has n complex components respectively received by antennas 120-1 through 120-n. Processor 60 further processes the preprocessed signal vectors to detect the m constituent data substreams. Symbol processor 65 processes the symbols to determine the data substreams (also herein "bit decisions") respectively cor-

responding to the symbols. Symbol processor 65 then stores the "decision bits" in memory 61 so that the decision bits may be used to cancel interference in further processing of the received signal vector. When all of the bits in a transmitted vector have been detected, then multiplexer 70 multiplexes the bits from the various substreams to form an estimate of the original data stream outputted by source 50 (FIG.1).

Interference Cancellation

[0027] For the following assume that receiver 200 receives a transmission burst of κ m-dimensional transmit vectors as impaired by the environment of channel 125 plus additive noise. (Note that the following discussion is given in the context of processing just one of the received vectors. It is to be understood of course that such processing/dectection equally pertains to each of the received vectors.) Each transmitted vector is received by n receiver antennas, e.g., 12 dimensional transmit vectors received via 16 antennas. Also consider that the decision statistics are to be stacked from the bottom up - i.e., as levels (1), (2),... (m) as shown in FIG. 3, in which the vector at the first (bottom) level offers the largest SNR for the first of the transmitted components that will be detected. For such an iteration of the vector signals, assume that receiver 200 has composed the first i-1 decision statistic vectors $d^{[\cdot]}$s for the levels up to the level designated "next" in FIG. 3 and that the i-1 decisions that are based on the decision statistics for (1), (2), (i-1) are free of errors. Such decisions may be used to cancel interference stemming from the components of q that have been already determined. Note that $q_{(j)}$, j = 1, 2, ... m, denotes the reordered components of q corresponding to levels (1), (2) .... (m). Also note that for the following discussion, it is useful to express h in terms of its m n-D columns so that (in discussing the formation of the decision statistic for detecting q(i) $h = [h_1\ h_2 \ldots h_m]$.

[0028] Also note that the received signal, r, is the n-D vector expressed as follows:

$$r = q_1 \cdot h_1 + q_2 \cdot h_2 + q_3 \cdot h_3 \ldots + q_m \cdot h_m + \nu \tag{5}$$

[0029] Note that each of the m $h_{(j)}$ may be defined using (5) as h_l (where: $1 \le l \le m$). From (5), $h_{(j)}$ is defined by the subscripted h that multiplies $q_{(j)}$ in (5). Also, r may be expressed as follows:

$$\begin{aligned} r = {} & [q_{(1)} \cdot h_{(1)} + q_{(2)} \cdot h_{(2)} + \ldots + q_{(i-1)} \cdot h_{(i-1)}] + q_{(i)} \cdot h_{(i)} \\ & + [q_{(i+1)} \cdot h_{(i+1)} + q_{(i+2)} \cdot h_{(i+2)} + \ldots + q_{(m)} \cdot h_{(m)}] + \nu \end{aligned} \tag{6}$$

[0030] Note, the first bracketed sum, $[q_{(1)} \cdot h_{(1)} + q_{(2)} \cdot h_{(2)} + \ldots + q_{(i-1)} \cdot h_{(i-1)}]$, is assumed to involve only correctly detected signal components and is subtracted from r to give the n-D vector $u^{[i]}$ defined by:

$$\begin{aligned} u^{[i]} = {} & [q_{(i)} \cdot h_{(i)}] + [q_{(i+1)} \cdot h_{(i+1)} + q_{(i+2)} \cdot h_{(i+2)} \\ & + \ldots + q_{(m)} \cdot h_{(m)}] + \nu \end{aligned} \tag{7}$$

[0031] Thus, in the processing of the same received vector r that is designated "next" in FIG. 3 (i.e., stack level 8), processor 60 cancels (subtracts) from vector r the vector $[q_{(1)} \cdot h_{(1)} + q_{(2)} \cdot h_{(2)} + \ldots + q_{(7)} \cdot h_{(7)}]$ as a result of having already determined/detected the latter vector. Processor 60 then "nulls" out of the vector being processed (e.g., the level 8 vector), the interference from transmitted signal components that have not yet been detected, i.e., the interference stemming from the transmission of $q_{(9)}$ through $q_{(12)}$, as shown in FIG. 3.

Interference Nulling Using Spatial Matched Filters

[0032] For those components (i+1) (i+2), ... (m) of q that have not yet been determined/detected, $u^{[i]}$ may be projected orthogonal to the m-i dimensional space spanned by $h_{(i+1)}, h_{(i+2)}, \ldots h_{(m)}$. The result of that projection is noted herein as $v^{[i]}$. The "interference nulling step"*, in a sense "frees" the detection process for $q_{(i)}$ of interference stemming from the simultaneous transmission of $q_{(i+1)}, q_{(i+2)}, \ldots q_{(m)}$.

The direction and magnitude of the spatial-matched-filter vector $d^{[i]}$ is considered next. Note that $q_{(i)}$ multiplies each component of $v^{[i]}$ so that the vector $v^{[i]}$ is similar to the situation of receiving an [n-(i-1)]-fold diversity interference-free signal in vector AWGN. Explicitly, the decision statistic for $q_{(i)}$ is the scalar product $< v^{[i]}, d^{[i]} >$, in which the noise power

* In the asymptote of high SNR the incremental advantage of not nulling, but instead maximizing the signal to noise plus self-interference is negligible.

of the scalar product is proportional to $\|d^{[i]}\|^2$. A standard result about optimum receive diversity may be used to represent the optimized signal to noise ratio, $SNR_{(i)}$. The resulting decision statistic also has the signal power that is proportional to $\|d^{[i]}\|^2$. It is thus convenient to define $\underline{v}^{[i]}$ to denote the vector $v^{[i]}$ in the hypothetical situation where space 125, FIG. 1, is free of additive noise. The $SNR_{(i)}$ of that medium is optimized when d[i] is any multiple of the value $\underline{v}^{[i]}$, as would follow by applying the well-known Cauchy-Schwarz inequality to the signal power term in the numerator in an expression for SNR. Indeed, the best of all of the opportunities for spatial filtering, is when the vector used in a scalar product to collapse $v^{[i]}$ to a scalar decision statistic is in the direction of $\underline{v}^{[i]}$: The reason for this is that when the collapsing vector is proportional to $\underline{v}^{[i]}$, the Cauchy-Schwarz upper bound on SNR is achieved with equality. Note that $\|\underline{v}^{[i]}\|^2$ appears multiplicatively in both numerator and the denominator of the optimum SNR, therefore, the optimum SNR is invariant with respect to the $\|\underline{v}^{[i]}\|$. While $d^{[i]}$ has the direction of $\underline{v}^{[i]}$ the scale of $d^{[i]}$ is simply set in accordance with the (arbitrary) scale factor used for the decision regions employed in the final stage of QAM detection. (Note that the processes of canceling, nulling and compensating is shown graphically in FIG. 4, as mentioned above.)

Compensation

[0033] Processor 60 upon receiving a burst of signal vectors stores the signal vectors in memory 61. As is done conventionally, the burst may contain information which receiver 200 may use to "learn" the transmission characteristics of the transmission environment 125 (FIG. 1). Such conventional learning (or training) information may be positioned at, for example, either the beginning (preamble) or mid point (midamble) of a burst of signal vectors, as is well-known.

[0034] After storing the received training vectors in memory, processor 60 then determines the stacking order in which the transmitted data symbols should be detected as a function of their respective SNRs. This is done to minimize the probability of making a detection error in the processing of a burst of received vectors. The determination includes forming spatially-matched-filter vectors by iteratively determining the order in which the transmitted symbols in a vector symbol are detected from a received vector. Note that the goal of such re-ordering is to maximize the minimum signal-to-noise ratio of the detection process. Processor 60, more specifically, stacks the m decision statistics for the m components in accordance with the following criterion:

$$\text{maximize minimum } [SNR_{(i)},\ 1 \le i \le m] \quad (8).$$

The reason that this criterion corresponds to minimizing the probability of a burst error is that in a high SNR situation (i.e., high p situation) the probability that a burst contains at least one error could be dominated by the $q_{(i)}$ that has the least $SNR_{(i)}$ (as will be shown below in connection with equations (10) and (11).

[0035] Processor builds the stack using a so-called "myopic" optimization procedure, which turns out to be the global optimization procedure expressed by equation (8) -- meaning that processor 60 starts at the bottom level of the stack and proceeds iteratively up to the m-th level, always choosing the next decision statistic among the options that maximizes the SNR for that level. With myopic optimization, processor 60 need only consider ~$m^2/2$ options in filling the totality of all m stack levels as opposed to a thorough evaluation of m! stacking options.

[0036] (As an aside, note that the improved compensation feature may be achieved using an improved iterative solution that is somewhat more computational. Specifically, assume that processor 60 is proceeding next to the i-th stack level. Once it makes the i-th decision, assume that no error was made. For that case, then, processor 60 may subtract the complex number corresponding to the constellation point from the decision statistic. This process provides processor 60 with the value of the noise that was included in the i-th decision statistic. This noise is correlated with corresponding additive noises of decision statistics further up the stack. Consequently, each additive noise term in subsequent decision statistics can be adjusted by subtracting off the conditional expectation of that noise conditioned on all the contributions of the assumed known additive noises of decision statistics lower in the stack.)

[0037] The program which implements the above described vertical ordering/layering in the processor 60 of FIG. 2 is shown in flow chart form in FIGs. 5 and 6. Specifically, FIG. 5 illustrates the way in which the optimum decision statistic vector $d^{[i]}$ for each of m stack levels, I = 1, 2, 3,..... m is determined. When the program is entered at block 500, following the storage of a received burst of signal vectors and the processing of the training information, the program proceeds to block 501 where it sets a variable i to a value of one and then proceeds to block 502. Processor 60, at block 502, processes the generic signal vectors to identify the spatial-matched-filter vector having the largest SNR of all of the candidates (as discussed above and as will be further discussed below). When processor 60 identifies that vector, it then (block 503) scales the vector relative to a respective scaling value. Processor 60 then stores the scaled vector in memory 61. Processor 60 then exits (blocks 505 and 506) if it determines that it has completed the forming and ordering (stacking) of all of the spatially-matched-filter vectors (i.e., i = m). If it has not completed the forming and ordering process, then processor 60 (blocks 505 and 507) increments i and returns to block 502 to find

which of the remaining candidates offers the largest SNR and places that candidate next in the stack.

[0038] An expanded view of block 502 is shown in FIG. 6. In particular, assume that processor 60 has already stored the channel matrix in memory 61. Also assume that processor 60 has already stacked in the preferred order i-1 the matching vector candidates and that processor 60 is now forming the remaining (m - (i - 1)) candidate vectors to determine which one of those candidates will be inserted at the ith level of the stack. At block 601, processor 60 forms a generic linear combination of the i-1 vectors positioned at the lower levels in the stack (and which have already been ordered in the stack), containing interferers. Processor 60 subtracts (block 603) that combination from the generic noise-free vector read (block 602) from memory 61, thereby leaving a representation of the vector for the transmitted signal components that have not yet been detected. Processor 60 then (block 608) initializes a variable j to a value of one to ensure that it will process all of the vector signals. For each vector candidate, e.g., the ith candidate vector, processor 60 (block 605) projects that vector orthogonally away from the (m- (i-1)) vector signals that interfere with the ith candidate vector to eliminate those interferers from the decision process. The resulting vector should be a vector that is free of interference from the other ones of the received vector signals. Processor 60 then (block 607) measures the norm (corresponding to the square of the length of the vector with respect to an origin in which each of the n components equals zero) to determine the value for that norm. The value of the norm is proportional to the SNR for that decision statistic. Processor 60 then determines whether the SNR is the best (largest) SNR for all of the candidates processed thus far. If it is, then processor 60 stores (block 609) the spatially-matched-filter vector having that SNR and its associated norm in the ith level of the stack and then proceeds to block 608 to continue the processing of the remaining candidates. If it is not, then processor 60 goes directly to block 608. After storing the selected signal vector candidate and its norm in memory, processor 60 then checks to see if it is done. If not, processor 60 increments j and proceeds to block 605, otherwise, it proceeds to block 503, FIG. 5.

[0039] Accordingly, then, processor 60 operating under control of the foregoing program places, for detection, the transmitted signal components in an optimum order based on their respective SNRs. As discussed above, processor 60 then uses each of the stacked spatially-matched-filter vectors to determine a bit combination/decision that a processed symbol most likely represents. As mentioned above, the procedure for determining a symbol is illustrated in FIG. 4, and a discussion of that procedure is repeated, but in the context of FIGs. 7 and 8.

[0040] Specifically, the program which implements in processor 60 (and somewhat in processor 65) the above nulling, canceling and matching steps 401, 402 and 403 illustrated in FIG. 4 is shown in flow chart form in FIGs. 7 and 8. Starting with FIG. 7, processor 60 begins such processing by entering the program of FIG. 7 at block 700. At that point, processor 60, initializes (block 701) a variable i to a value of 1 and uses to point to a respective level in the stack. For the following discussion, assume that processor 60 is processing the ith level in the stack. Processor 60 (block 702) processes the vector signal (as shown for procedures 401, 402 and 403, FIG. 4) positioned at the ith level in the stack to determine the symbols $q_{(i)}$ most likely represented by the signal vector. Processor 60 stores the bit decisions corresponding to $q_{(i)}$ in memory 61, and then checks (block 703) the value of i to see if it has completed m successive processings of the received vector signal r. If so, then processor 60 exits the program. Otherwise, processor 60 (block 705) increments i to point to the next succeeding level in the stack and proceeds to block 702 to process the received signal vector in a manner corresponding to that level.

[0041] An expanded version of block 702 is shown in FIG. 8. Specifically, at block 801 processor 60 reads from memory 61 the received n-dimensional vector signal r(t), and then reads (block 802) from memory 61 those signal vector components that have already been processed/detected in accordance with block 702, i.e., those components below the ith level in the stack. Then, in the manner discussed above, processor 60 cancels the i-1 retrieved signal vectors, $q_{(1)}\cdot h_{(1)}$, $q_{(2)}\cdot h_{(2)}, \ldots, q_{(i-1)}\cdot h_{(i-1)}$, out of r and then proceeds to block 803 with a signal vector substantially devoid of interference from those i-1 vectors. Processor 60 then projects the resulting vector away from the interferers that have not yet been detected. That is, processor 60 reads from memory the spatial match vector $d^{[i]}$ determined at block 504 for the ith level of the stack. Processor 60 (cooperating with processor 65) then takes the scalar product of the spatial match vector and the result generated at block 803 to generate a complex number.

[0042] Processor 65 (805) then determines which one of a conventional multipoint signal constellation, e.g., a sixteen point signal constellation (as shown for constellation 404, FIG. 4), is closest to the complex number. An example of such a point is shown in FIG. 4 in which point 405 representing the complex number is closest to the constellation point of quadrant 406. Processor 65 (block 806) then stores the data bit decisions represented by the identified constellation point in memory and then returns control to processor 60. Processor 60 then proceeds to block 703 to process the next level in the stack.

[0043] When the received signal vector has been so processed and all symbols detected, with the corresponding bits decisions stored in memory, then multiplexer 70 multiplexes the bit decisions to an output terminal.

[0044] The advantages of the foregoing may be appreciated by way of an example based on a particular set of parameters, in which, for example; ρ = 18 dB, and (m,16) with $m \leq 16$, and in which transmitter 100 and receiver 200 (FIG. 1) may have antenna arrays of up to 16 antennas. Also, assume that 95% of the transmission bursts are free of errors (i.e., at most a 5% outage), and an ideal Rayleigh propagation environment for space 125. Further assume that

each burst contains, besides training vectors, 100 vector symbols. For the assumed parameters and 16 x 16 system, the Shannon capacity is 79. 1 bps/Hz. (Note that "Shannon capacity" is a well-known term in the art.)

[0045] The number of transmit antennas, m, and the number of points in each of the planar constellations, K, may be optimized to maximize the number of bps/Hz. The number of constellation points in m-D (or n-D) complex space may be expressed as follows:

$$\text{Number of constellation points} = [\text{Number of 2-D constellation points}]^{[\text{Number of substreams}]} = K^m \qquad (9)$$

The optimization process involves performing the following procedure to iteratively explore m = 1, 2, 3, ... 16. For each of these cases, as many bits per 2-D constellation may be used, to the point where one more bit would violate the so-called 5% "outage constraint"[1].

The equation for the probability of at least one error in a block with K vector symbols is shown below as equation (10). Evaluation of this probability requires that the $SNR_{(i)}$ for $1 \le i \le m$. Monte-Carlo generated H realizations may be used to get samples of the m SNRs. For K point QAM constellations the formula in the large ρ realm is

$$\text{Prob[Erroneous Block]} \approx \kappa \times \sum_{i=1}^{m} P_b(SNR_{(i)}) \qquad (10)$$

where $P_b(.)$ is the well known function for the probability of bit error of a 2-D constellation as a function of SNR, namely:

$$P_b(SNR_{(i)}) \approx [(K^{½} - 1)/(K^{½} \cdot \log_2 K)] \times p^{-½} \cdot \int_a^{\infty} \exp(-x^2)\,dx \qquad (11)$$

where $a = [(3 \cdot SNR_{(i)})/(2 \cdot (K - 1))]^{½}$.

[0046] Using equations (9) and (10), and starting with m = 1 the system of FIG. 1 may support K = 128 point constellations or equivalently 7 bps/Hz. For m = 2 the system can support a 32 point planar constellation of 5 bps/Hz. For m = 7, the system can support 16 point constellations of 4 bps/Hz. For m = 12, the system can support 3 bps/Hz, which is one of the higher dimensional constellations of 8^{12} = 68,719,476,736 points or 36 bps/Hz.

[0047] The foregoing is merely illustrative of the principles of the invention. Those skilled in the art will be able to devise numerous arrangements, which, although not explicitly shown or described herein, nevertheless embody those principles that are within the spirit and scope of the invention. For example, instead of maintaining a constant BER and outage probability, the skilled artisan could maintain a constant transmitted power, and express the relative merit of the two approaches in terms of a difference in outage probability as a function of rate. As another example, the skilled artisan could hold the BER and outage probability constant for both systems, and express the relative merit in terms of transmitter power, or, the life of a battery in a portable system.

[0048] In addition, those skilled in the prior art would recognize from the foregoing that the notion of using more transmit antennas than transmit radios at the transmitter and selecting a subset of antennas on which to transmit, could be similarly applied to the receiver. In that case, more receive antennas would be deployed than receive radios and a subset of the receive antennas on which to receive a transmitted vector signal would be selected.

[0049] Those skilled in the relevant art would also recognize from the foregoing that in certain applications of the claimed invention it may be desirable to use only a subset of the inventive features. For example, it may be desirable to use just nulling but not cancellation and reordering, or else nulling and cancellation but not ordering.

Claims

1. A communications system comprising

a transmitter having k antennas, said transmitter responsive to receipt of an m-dimensional transmit symbol

[1] When the number of constellation points was greater than two and not a perfect square we used a regular constellation with good distance properties. For example, for an 8 point constellation we used a square with four equilateral triangles attached to the four sides and pointing outward. The vertices of the square, along with the four triangle vertices that oppose each of the four sides of the square made up the constellation

vector from a source, components of said transmit symbol vector comprising QAM symbols, said transmit symbol vector being transmitted over m of the k antennas using a predetermined modulation technique, where $k \geq m > 1$, and CHARACTERIZED BY

a receiver having n antennas for receiving signals from said transmitter as n-dimensional received signal vectors, where $n \geq m$, each of said received signal vectors comprising a linear combination of symbols from said transmitter and additive noise,

wherein said receiver further comprises

a detection processor that processes the n-dimensional received signal vector to form an estimate of the m-dimensional transmit symbol vector, the detector further comprises

a processor that (a) determines a preferred permutation of integers 1, 2, ... m, which define an order in which said m components of said transmit symbol vector are estimated, and in which the preferred permutation is a function of the signal-to-noise ratios of the m components, and

(b) in the order defined by said preferred permutation then estimates the (i)-th ordered component of the transmit symbol vector by nulling out from the received signal vector contributions due to transmit symbol vector components (i+1), (i+2),..... (m) which have not yet been estimated, and canceling out from the received signal vector contributions due to transmit symbol vector components (1), (2),... (i-1) which have already been estimated, where (i) denotes the ith element of the preferred permutation.

2. The communications system of claim 1 wherein the detection processor is arranged to repeatedly process received training signals characterizing the signal propagation environment to generate a set of m spatially matched filter vectors, offering the best Signal-to-Noise Ratio (SNR) for detecting the m transmitted symbols.

3. The communications system of claim 1 wherein the order for detecting the m transmitted symbols maximizes the minimum of the m signal-to-noise ratios of the detection process.

4. The communications system of claim 1 wherein the selection of the transmitter antennas is randomly changed prior to the transmission of a group of transmit vector symbols.

5. The communications system of claim 1 further comprising a feedback channel from the receiver to the transmitter and wherein the selection of the transmitter antennas is optimized based on signal propagation environment information that the receiver supplies to the transmitter via the feedback channel.

6. The communications system of claim 1 further comprising a feedback channel from the receiver to the transmitter and wherein the transmitter transmits substreams of a demultiplexed stream of symbols supplied by a source over respective ones of a predetermined set of transmit antennas and changes the selection of antennas forming the set of antennas based on signal propagation environment information supplied by the receiver via the feedback channel.

7. The communications system of claim 1 further comprising a feedback channel from the receiver to the transmitter and wherein the transmitter transmits a vector symbol over a subset of the k transmitter antennas, in which the subset is selected based on signal propagation environment information supplied by the receiver via a feedback channel.

8. The communications system of claim 1 wherein the receiver has an arbitrary number of receive antennas greater than n and wherein the n antennas that are used to receive signals from said transmitter as n-dimensional received signal vectors is a subset of the arbitrary number of receive antennas.

9. The communications system of claim 1 wherein the rate at which symbols may be received accurately at the receiver is proportional to the number of transmit antennas used to transmit the symbols and logarithmic in the level of transmitted power so that the level of power at which symbols may be transmitted at the transmitter may be substantially decreased by increasing the number of transmit antennas by a relatively small number.

10. A wireless transmitter comprising

a source of a stream of symbols,

a plurality of transmitter antennas, and CHARACTERIZED BY

a transmitter processor which demultiplexes the symbol stream into m substreams of symbols and which then transmits each substream of symbols over a selected one of the transmitter antennas using a predetermined

modulation technique.

11. The transmitter of claim 10 wherein said plurality of transmitter antennas is greater than m antennas, where m > 1, and wherein selection of the m transmitter antennas used to transmit the m substreams of symbols is arbitrary.

12. The transmitter of claim 10 wherein the selection of the transmitter antennas is randomly changed prior to the transmission of a group of symbols.

13. The transmitter of claim 10 further comprising

    a receiver for receiving the transmitted symbols, and
    a feedback channel from the receiver to the transmitter and wherein the transmitter processor optimizes the selection of the transmitter antennas based on signal propagation environment information that the receiver supplies to the transmitter via the feedback channel.

14. The system of claim 10 further comprising a feedback channel from a receiver of the transmitted symbols to the transmitter and wherein the transmitter transmits a vector symbol over a subset of the plurality of transmitter antennas, in which the subset is selected based on signal propagation environment information received from the receiver via the feedback channel.

15. A wireless receiver comprising

    a plurality of receiver antennas for respectively receiving a plurality of n signal components forming a received signal vector, where n > 1, and CHARACTERIZED BY
    a processor that stores a received signal vector in memory with other received signal vectors forming a burst of signal vectors, and
    a detection processor which processes the stored received signal vector to determine components of a transmitted signal vector in an order determined as a function of respective signal-to-noise ratios determined for particular decision statistics so that (a) interference stemming from transmitted symbol vector components that have been processed are canceled out of a signal vector that is currently being processed, (b) interference stemming from transmitted symbol vector components that have not yet been processed is nulled out of the signal vector that is currently being processed by projecting the signal vector orthogonal to a space occupied by the latter interference, and then processing the projected signal vector in accordance with a predetermined demodulation technique to identify the components of the transmitted symbol vector.

16. The receiver of claim 15 wherein the detection processor includes repeatedly processing data characterizing the transmission environment to generate a set of m spatially matched filter vectors offering the best Signal-to-Noise Ratio (SNR) for detecting the components of the transmitted symbol vector.

17. The receiver of claim 15 wherein the order for detecting the components of the transmitted symbol vector is to maximize the minimum signal-to-noise ratio of the detection process.

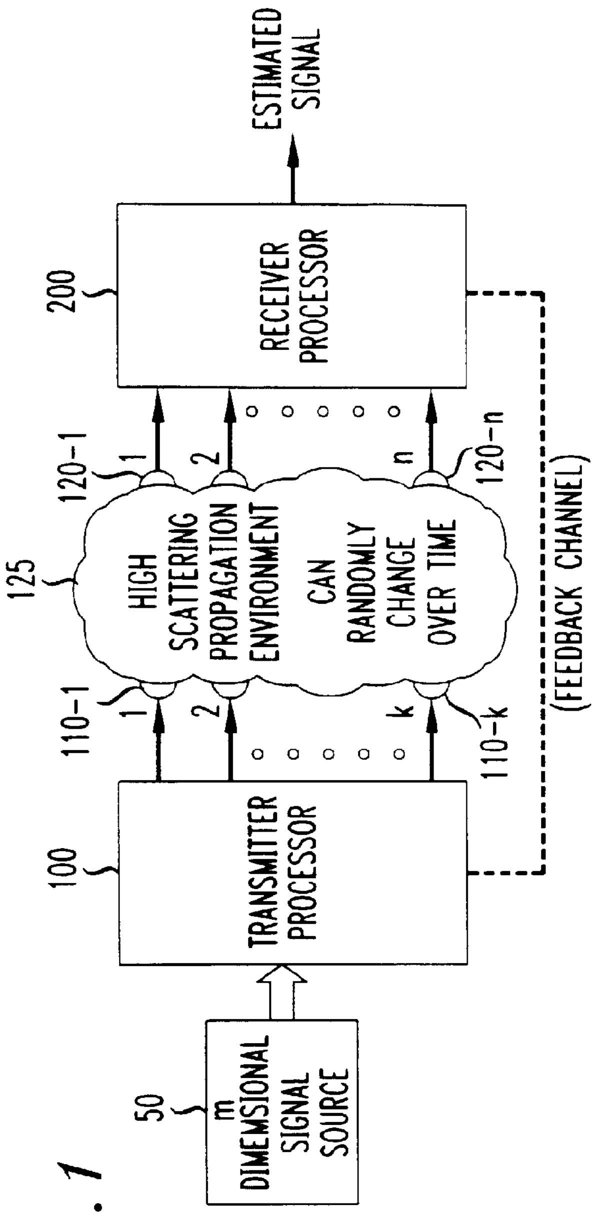


FIG. 1

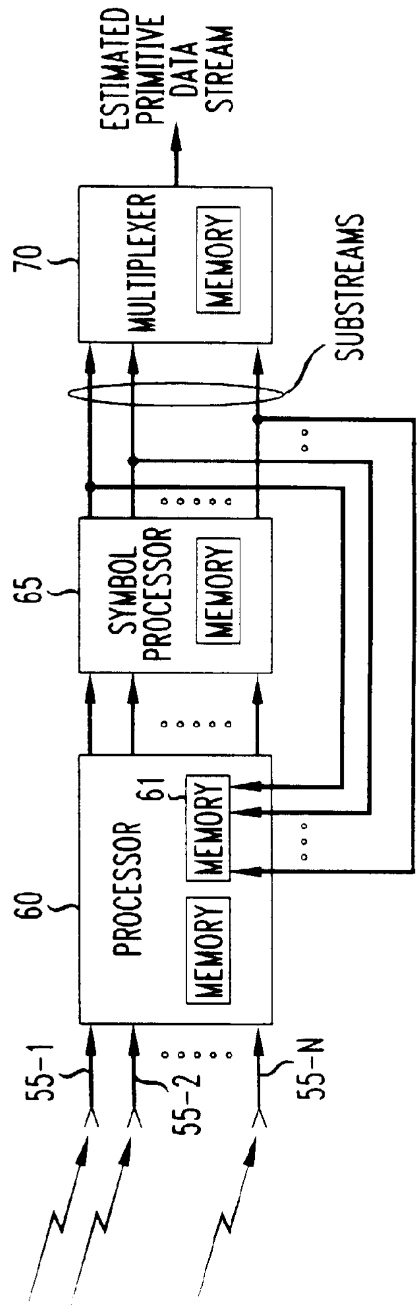


FIG. 2
200

FIG. 3

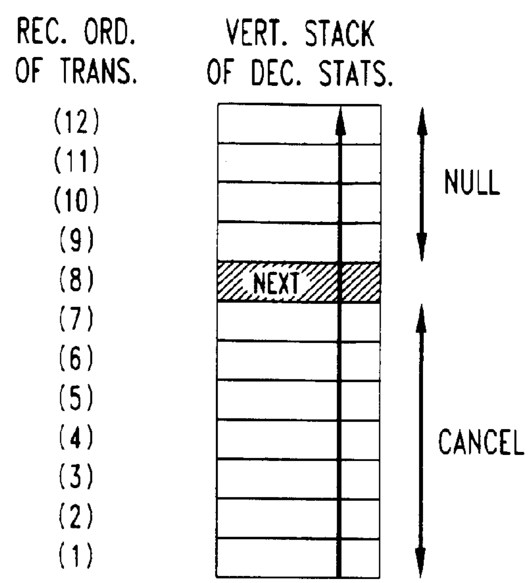

REC. ORD.
OF TRANS.
VERT. STACK
OF DEC. STATS.
(12)
(11)
(10)
(9)
(8)
(7)
(6)
(5)
(4)
(3)
(2)
(1)
NEXT
NULL
CANCEL

FIG. 4

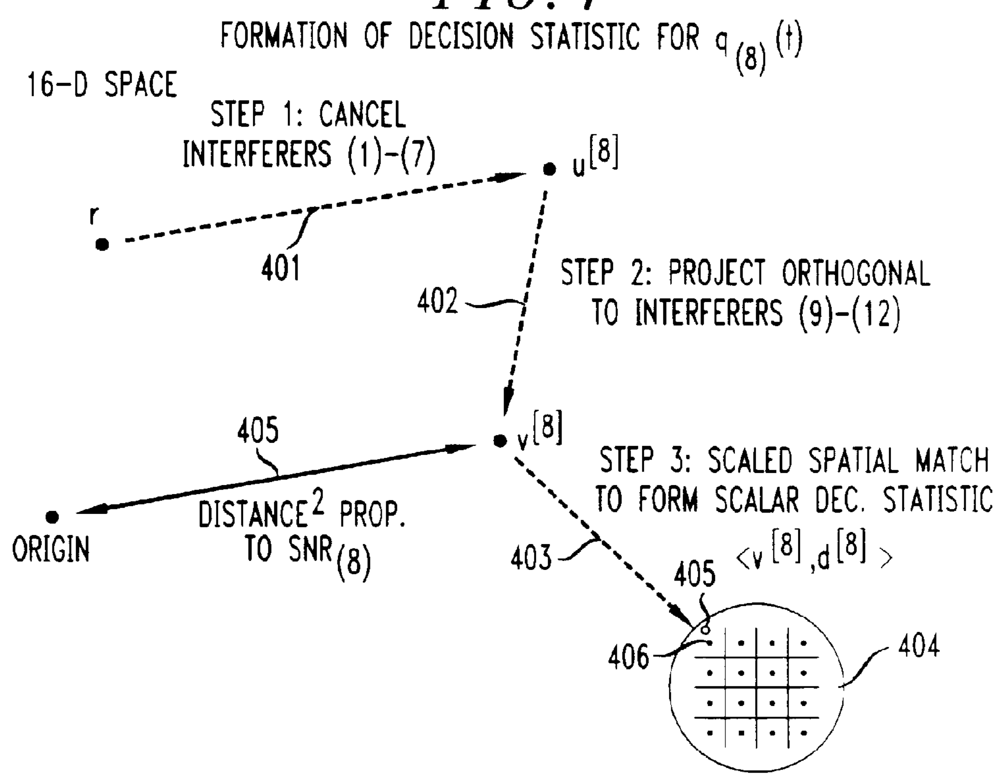

FORMATION OF DECISION STATISTIC FOR $q_{(8)}(t)$
16-D SPACE
STEP 1: CANCEL
INTERFERERS (1)-(7)
$u^{[8]}$
r
401
402
STEP 2: PROJECT ORTHOGONAL
TO INTERFERERS (9)-(12)
405
$v^{[8]}$
STEP 3: SCALED SPATIAL MATCH
TO FORM SCALAR DEC. STATISTIC
$\langle v^{[8]}, d^{[8]} \rangle$
ORIGIN
DISTANCE^2 PROP.
TO $SNR_{(8)}$
403
405
406
404

FIG. 5

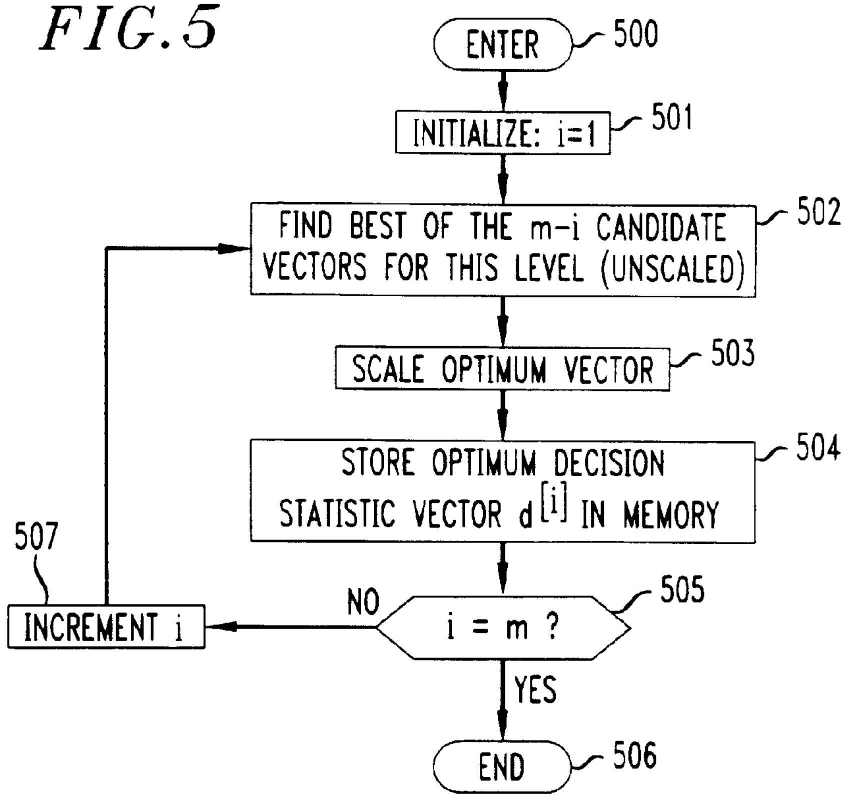

ENTER
500
INITIALIZE: i=1
501
FIND BEST OF THE m-i CANDIDATE VECTORS FOR THIS LEVEL (UNSCALED)
502
SCALE OPTIMUM VECTOR
503
STORE OPTIMUM DECISION STATISTIC VECTOR d[i] IN MEMORY
504
i = m ?
505
NO
INCREMENT i
507
YES
END
506

FIG. 7

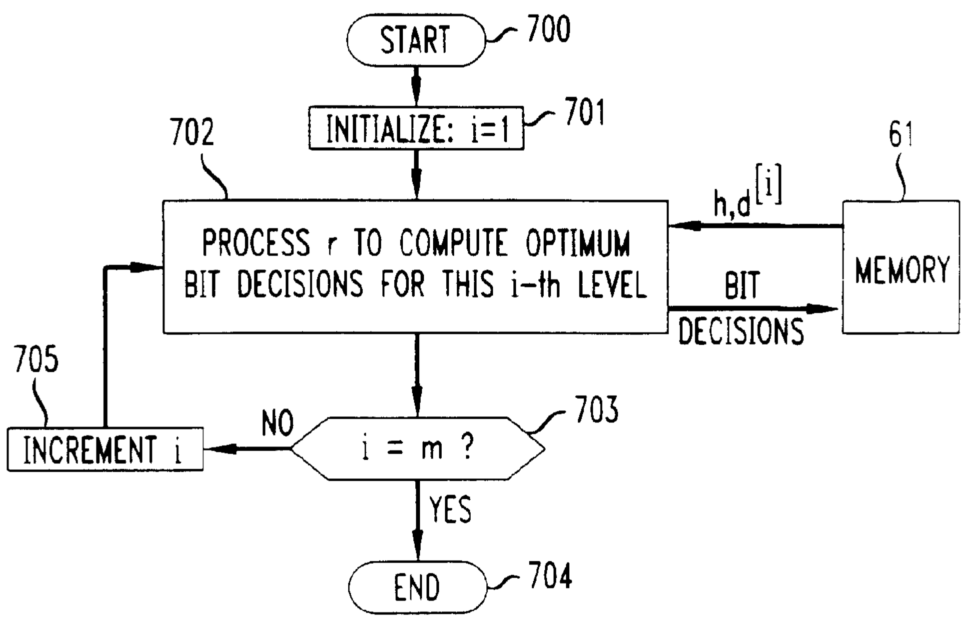

START
700
INITIALIZE: i=1
701
702
PROCESS r TO COMPUTE OPTIMUM BIT DECISIONS FOR THIS i-th LEVEL
h,d[i]
BIT DECISIONS
MEMORY
61
i = m ?
703
NO
INCREMENT i
705
YES
END
704

FIG. 6

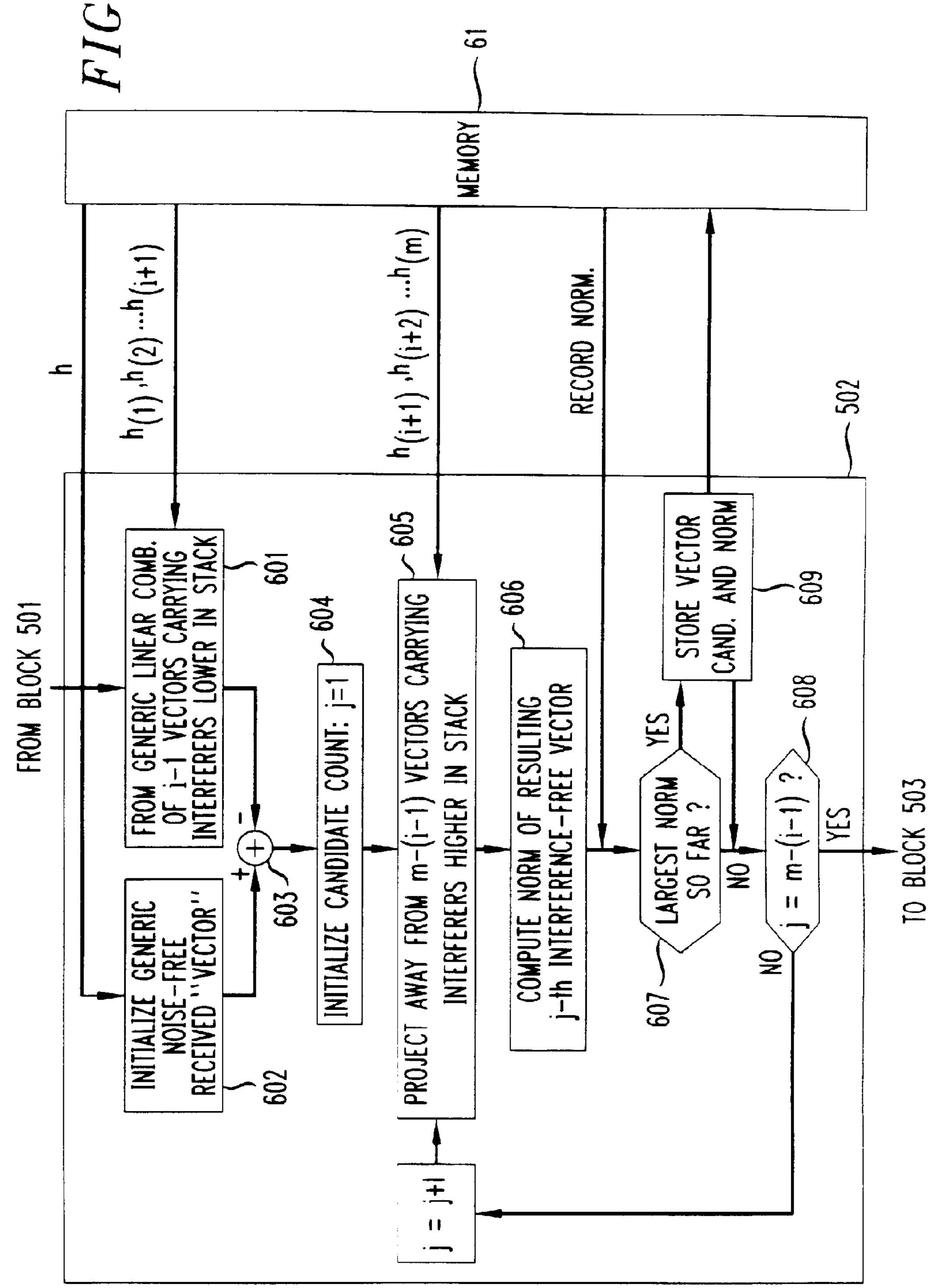
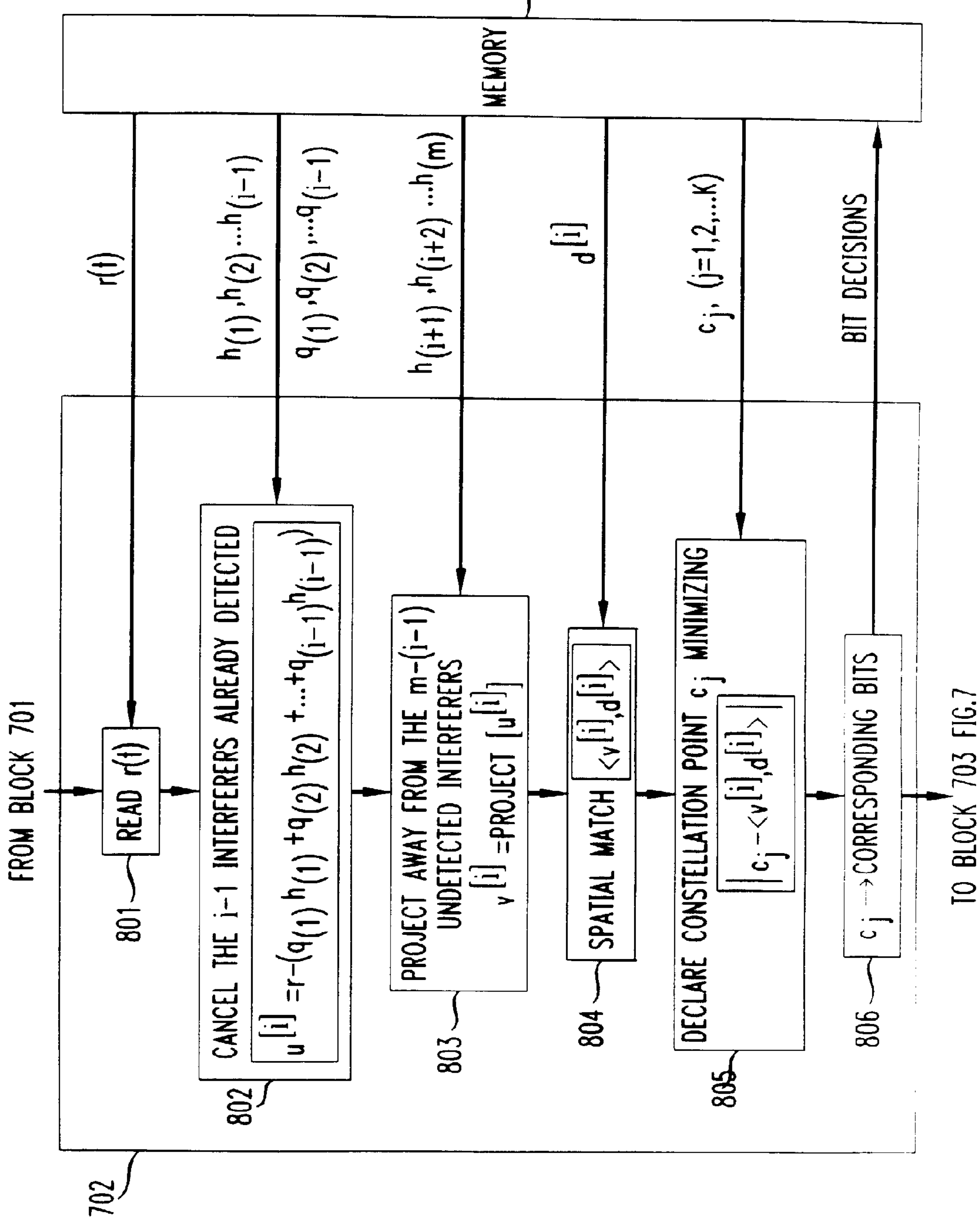

61
MEMORY
h
h(1), h(2) ...h(i+1)
h(i+1), h(i+2) ...h(m)
RECORD NORM.
502
FROM BLOCK 501
FROM GENERIC LINEAR COMB. OF i-1 VECTORS CARRYING INTERFERERS LOWER IN STACK
601
INITIALIZE GENERIC NOISE-FREE RECEIVED "VECTOR"
602
+
-
603
INITIALIZE CANDIDATE COUNT: j=1
604
PROJECT AWAY FROM m-(i-1) VECTORS CARRYING INTERFERERS HIGHER IN STACK
605
COMPUTE NORM OF RESULTING j-th INTERFERENCE-FREE VECTOR
606
LARGEST NORM SO FAR ?
607
YES
NO
STORE VECTOR CAND. AND NORM
609
j = m-(i-1) ?
608
YES
NO
TO BLOCK 503
j = j+1

FIG.8


MEMORY
61
r(t)
$h_{(1)}, h_{(2)} \ldots h_{(i-1)}$
$q_{(1)}, q_{(2)}, \ldots q_{(i-1)}$
$h_{(i+1)}, h_{(i+2)} \ldots h_{(m)}$
$d^{[i]}$
c_j, (j=1,2,...K)
BIT DECISIONS
FROM BLOCK 701
READ r(t)
801
CANCEL THE i-1 INTERFERERS ALREADY DETECTED
$u^{[i]}=r-(q_{(1)}h_{(1)}+q_{(2)}h_{(2)}+\ldots+q_{(i-1)}h_{(i-1)})$
802
PROJECT AWAY FROM THE m-(i-1) UNDETECTED INTERFERERS
$v^{[i]}$=PROJECT $[u^{[i]}]$
803
SPATIAL MATCH $\langle v^{[i]},d^{[i]}\rangle$
804
DECLARE CONSTELLATION POINT c_j MINIMIZING
$\left|c_j-\langle v^{[i]},d^{[i]}\rangle\right|$
805
$c_j \rightarrow$ CORRESPONDING BITS
806
TO BLOCK 703 FIG.7
702

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(71) Applicant:
SUN MICROSYSTEMS, INC.
Palo Alto, California 94303 (US)

(72) Inventors:
- **Johnson, Thomas V.**
Burlington, MA 01803 (US)
- **Davis, James D.**
Pepperell, MA 01463 (US)
- **Ting, Charles C.**
Wayland, MA 07118 (US)

(74) Representative:
Browne, Robin Forsythe, Dr.
Urquhart-Dykes & Lord
Tower House
Merrion Way
Leeds LS2 8PA West Yorkshire (GB)

(54) **Web-based enterprise management with multiple repository capability**

(57) A transport neutral technique allows an object manager (20) to communicate with a CIM repository (130) using any of a variety of protocols (158, 160). The object manager software (20) is independent of the transport mechanism used and need not be changed if the transport mechanism changes. A computer system (110) to be managed includes a CIM object manager (20) and any number of provider APIs (122) that provide resource information about the computer system. A CIM repository (130) stores classes and instances used by the object manager. A remote application computer (150) runs a software management application (32) that communicates with the object manager of the computer system using a local client API (156). A Repository API of the object manager includes an interface definition (300) defining all methods called by the object manager. Also included is a protocol-specific class that implements the interface definition; there is a protocol-specific class for each protocol desired to be supported. Each class implements methods using a specific protocol. A factory class is executed when the object manager invokes a method call passing in a desired protocol parameter. The factory class creates a protocol-specific object of one of the protocol-specific classes depending on the protocol parameter. The object is returned to the object manager which executes one of its protocol-specific methods thus allowing communication to a repository using a protocol independent of the object manager.

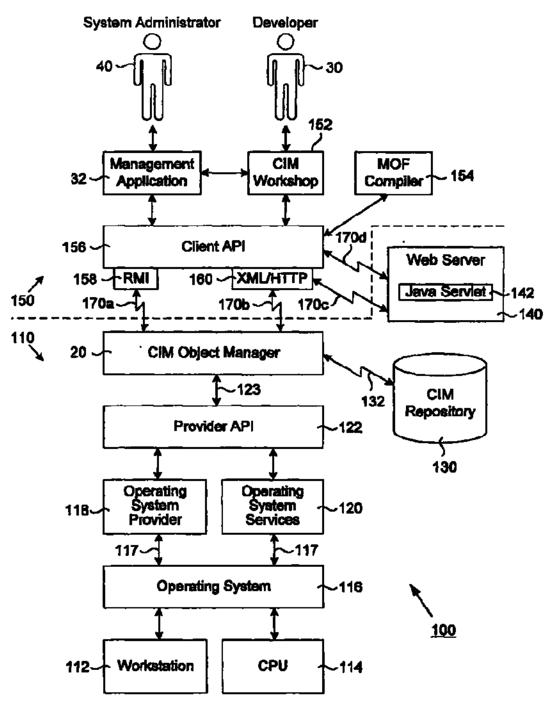

FIG. 2A

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Description

FIELD OF THE INVENTION

[0001] The present invention relates generally to managing the resources of a computer system. More specifically, the present invention relates to a technique for communicating database operations from a Common Information Model (CIM) object manager to multiple remote CIM repositories.

BACKGROUND OF THE INVENTION

[0002] Recently, computers and their associated peripheral equipment (a computer system) have become increasing more complex. As such, it has become progressively more and more complicated for a user or system administrator to manage the resources of such a computer system. With a variety of peripheral devices and software applications available for use, and their ever-changing nature, the job of a system administrator has become more difficult. Computer system resources such as attached devices, network connections, software applications, etc., must all be managed to ensure an efficiently working system for the user. Within a large corporation having large numbers of such computer systems spread around the world, the task of managing the resources of each computer system can be daunting.

[0003] Recently, the industry has responded to such a need by introducing Web-Based Enterprises Management (WBEM) which is both an initiative and a technology. As an initiative, WBEM includes a standard for managing systems, networks, users, and applications by using Internet technology. As a technology, WBEM provides a way for management applications to share management data independently of vendor, protocol, operating system, or management standard. By developing management applications according to WBEM principles, vendors can develop products that work together easily at a lower cost of development.

[0004] One known standard for implementation of WBEM is the Common Information Model (CIM). CIM is an approach to managing systems and networks. CIM provides a common conceptual framework to classify and define the parts of a network environment and depict how they integrate. The model captures notions that are applicable to all areas of management, independent of technology implementation.

[0005] WBEM software includes tools and technology that software developers can use to create CIM-compliant software applications that manage the environment of a computer system. Developers can also use this software to write "providers," programs that supply data and events for managed objects that are specific to their domain.

[0006] There can be drawbacks, however, associated with various implementations of WBEM software. For example, it may be necessary for the object manager of a computer system to be able to access different types of databases, whether local or remote. Not all implementations, however, are well-suited for this type of access.

[0007] FIG. 1 illustrates a prior art computer system 10 that has resources to be managed. Resources include disk usage, CPU utilization, running applications, etc. Not shown for simplicity are hardware components of the computer system or other software applications. A CIM object manager 20 is responsible for handling all communication between management applications, a CIM repository 26 and managed objects. In this example, CIM repository 26 is a local drive that communicates via a local connection 28 to object manager 20. A provider application programming interface (API) 22 provides an interface to any needed system information 24 to be provided to object manager 20. Object manager 20 may also access CIM repository 26 over local connection 28 to quickly retrieve data objects that have been previously stored. In this fashion, object manager 20 is well-suited for gathering resource information regarding computer system 10.

[0008] In order to efficiently manage the resources of computer system 10, a software developer 30 writes management application software 32 for managing the resources of the computer system. When in operation, the results of management application 32 may be used by a system administrator 40 to manage the computer system. Client application 32 communicates via a Client API 34 to retrieve resource information from computer system 10. Client API 34 uses any suitable local or remote network connection 36 to access object manager 20.

[0009] Prior art implementations of this sort use a single protocol for communication from object manager 20 to CIM repository 26 over local connection 28. Such an implementation is inflexible in that the object manager commands to repository 26 are dependent upon a single protocol. In other words, the commands are not independent of the protocol; should the protocol be modified or if another protocol be used or desired, it will be necessary to rewrite portions of object manager 20 which would be undesirable.

[0010] In addition, having an object manager that is depended upon a particular protocol presents difficulties when repository 26 is remote from object manager 20. In this scenario, it may be desirable to communicate over a network connection using any of a variety of protocols instead of always being required to use a local protocol. In prior art computer system 10 portions of object manager 20 would have to be rewritten for each and every different protocol that is desired to be used.

[0011] Therefore, a technique is desired that would permit an object manager to communicate both locally and remotely with any number of repositories using any of a variety of protocols. It is desired to implement this technique with the least impact upon developers of

object manager software.

SUMMARY OF THE INVENTION

[0012] To achieve the foregoing, and in accordance with the purpose of the present invention, a technique is disclosed that allows an object manager to communicate with any number of repositories using any of a variety of local or remote protocols. Advantageously, the object manager becomes independent of protocol used and need not be changed if the protocol changes.

[0013] In one embodiment, a method is used for communication between a Common Information Model (CIM) object manager and a CIM repository. The method involves first creating a connection between the object manager the CIM repository. Next, a protocol indicator is passed from the object manager to a repository API. The protocol indicator identifies a protocol by which the object manager desires to communicate with the CIM repository. A protocol-specific object is created having methods implemented using the protocol. Finally, the protocol-specific object is returned to the object manager, thus the object manager may communicate with the CIM repository using the protocol desired.

[0014] In another embodiment, a computer system interacts with a CIM repository on a separate computer. The computer system includes an object manager that has program code for interacting with the CIM repository and a protocol indicator. Also included is a repository application programming interface (repository API) that has a factory class arranged to receive the protocol indicator from the object manager and produce a protocol-specific object. Also within the repository API is a first class having methods defined thereon implemented in a first protocol and a second class having methods defined thereon implemented in a second protocol. Thus, the protocol-specific object may be returned to the object manager for use in communicating with the CIM repository using a desired protocol.

BRIEF DESCRIPTION OF THE DRAWINGS

[0015] The invention, together with further advantages thereof, may best be understood by reference to the following description taken in conjunction with the accompanying drawings in which:

FIG. 1 illustrates a prior art computer system that has managed resources.

FIGS. 2A and 2B illustrate a Web-Based Enterprise Management (WBEM) architecture suitable for implementing an embodiment of the invention.

FIGS. 3A and 3B illustrate an example graphical user interface for the CIM workshop of FIG. 2.

FIG. 4 illustrates an interface definition useful for implementing an embodiment of the invention.

FIG. 5 illustrates an implementation definition for the interface of FIG. 4.

FIG. 6 illustrates another implementation definition for the interface of FIG. 4.

FIG. 7 is a JAVA factory class definition.

FIG. 8 is a flowchart illustrating a store or retrieve repository method issued by an object manager to a CIM repository.

FIGS. 9 and 10 illustrate a computer system suitable for implementing embodiments of the present invention.

DETAILED DESCRIPTION OF THE INVENTION

[0016] FIGS. 2A and 2B illustrate a Web-Based Enterprise Management (WBEM) architecture suitable for implementing an embodiment of the invention. Architecture 100 includes computer system 110 having resources to be managed, and application computer 150 where management applications are developed and run. Computer system 110 may be any suitable computer having resources needing to be managed such as CPU load, disk space, installed applications, etc. The hardware layer includes workstation 112 and CPU 114. Workstation 112 may be any suitable computer workstation such as the SPARC workstation from Sun Microsystems, Inc. CPU 114 may be any suitable processor such as an Intel processor. The software layer of computer system 110 includes operating system 116 (other software applications not shown for simplicity) which may be any suitable operating system such as the Solaris operating system from Sun Microsystems, Inc.

[0017] The object provider layer of computer system 110 includes operating system provider 118, operating system services 120 and provider application programming interface (API) 122. The provider layer communicates with operating system 116 via interface 117, for example, by using a JAVA Native Interface (JNI). Object providers act as intermediaries between CIM object manager 20 and one or more managed devices. When object manager 20 receives a request from a management application 32 for data that is not available in CIM repository 130 it forwards the request to a provider. Object providers are installed on the same machine as object manager 20. Object manager 20 then uses provider API 122 to communicate with locally installed providers. Providers are classes that perform various functions in response to a request from object manager 20. For example, providers map information from a managed device to a CIM JAVA class and map

information from a CIM JAVA class to a managed device format.

[0018] Provider API 122 is an API used by various provider programs to communicate information about managed objects to object manager 20. Operating system provider 118 is a collection of JAVA classes or native methods that represent the operating environment of computer system 110. Operating system services 120 also provide logging information from operating system 116 to object manager 20.

[0019] A variety of other providers may also be present. For example, an SNMP provider includes JAVA classes that map CIM data to SNMP data. Also, a CPU-specific provider may be used to transport resource information directly between CPU 114 and provider API 122, thus bypassing operating system 116.

[0020] Providers may be categorized into three types according to the requests they service. An instance type supplies dynamic instances of a given class and supports the retrieval, enumeration, modification and deletion operations. A property type supplies dynamic property values, for example, disk space. A method type supplies methods of one or more classes. A single provider can support both methods and instances. Most providers are "pull" providers which mean they maintain their own data, generating it dynamical if necessary. Pull providers have minimal interaction with object manager 20 and CIM repository 130. The data managed by a pull provider typically changes frequently, requiring the provider to either generate the data dynamically or retrieve it from a local cache whenever an application issues a request. A single provider can act simultaneously as a class instance and method provider by proper registration and implementation of all relevant methods.

[0021] The management layer of computer system 110 includes CIM object manager 20, CIM repository 130 and web server software 140. Object manager 20 may be any suitable WBEM compliant manager. Object manager 20 manages CIM objects that are represented internally as JAVA classes. Client computers running management applications (such as application computer 150) connect to object manager 20 for resource information about computer system 110. When a WBEM client connects to object manager 20 it receives a reference to that object manager. The client can then perform WBEM operations using this reference.

[0022] When management application 32 uses client API 156 to request or update information about a managed object, object manager 20 contacts either the appropriate provider for that object or a suitable persistent storage mechanism such as repository 130. In one embodiment, classes that are handled by a provider have a "provider" qualifier that identifies the provider to contact for the class. When object manager 20 receives a request for a class that has a "provider" qualifier, it routes the request to the specified provider. If no provider is specified it routes the request to repository 130 using JAVA Naming and Directory Interface (JNDI) 132.

[0023] Object manager 20 also performs various start-up functions: starting and registering the RMI server; registering the XML server; setting up a connection to repository 130; and waiting for incoming requests. Object manager 20 also performs other normal operations: performing security checks such as authentication and authorization; performing syntactical and semantic checking of CIM data operations; routing requests to providers or persistent storage; and delivering data from providers or from persistent storage to client management applications.

[0024] CIM repository 130 is a central storage area for CIM class and instance definitions that communicates with object manager 20 via connection 132. Connection 132 may be any suitable local connection within computer system 110, or may be a remote connection. Connection 132 may use any suitable protocol. Further details on communication with repository 130 are provided in FIG. 2B and in FIGS. 4-8.

[0025] Web server software 140 may be any suitable WBEM XML-compliant web server such as the Sun Web Server available from Sun Microsystems, Inc. JAVA servlet 142 converts XML data to the client API format. For example, if management application 132 contains XML data, client API client 156 encodes the data as XML messages and transports the encoded messages to web server 140 that is running JAVA servlet 142. Web server 140 listens for XML messages on a standard port and passes control to servlet 142 when detected. Servlet 142 then decodes the XML messages it receives. Servlet 142 then converts the XML data to the client API format and transmits the information back to client API 156 in RMI format. Alternatively, should object manager 20 support the HTTP format, client API 156 may communicate directly to object manager 20 without the need for web server 140.

[0026] The application layer of WBEM architecture 100 includes management application 32, CIM workshop 152, MOF compiler 154 and client application programming interface (API) 156. In this embodiment of the invention, these elements of the application layer are shown running on application computer 150 (other hardware and software not shown for simplicity). Alternatively, application computer 150 and computer system 110 may be the same computer or the elements of the application layer may reside on a variety of computers and not exclusively on application computer 150.

[0027] A software developer 30 uses any suitable software tool to develop a management application 32 for processing and displaying data from managed objects of computer system 110. Management application 32 uses client API 156 to request information about managed objects from object manager 20. In this fashion, analysis of the resources of computer system 110 can be presented to a system administrator 40 for proper action.

[0028] Client API 156 and provider APIs represent

and manipulate CIM objects. These APIs represent CIM objects as JAVA classes. An object is a computer representation or model of a managed resource of computer system 110 such as a printer, disk drive or CPU. A developer uses the CIM specification to describe managed objects and to retrieve information about managed objects in computer system 110. One advantage of modeling managed resources using CIM is that those objects can be shared across any system that is CIM compliant.

[0029] Management application 32 may be any of a wide variety of software applications written to analyze and manage the resources of computer system 110. By way of example, management application 32 manages system aspects such as disk information (space available, partitions, etc.), CPU load, event processing, date, time, time zone, memory available, ports, etc.

[0030] Application 32 may also manage specific devices of the computer system such as disks, tape drives, modems, other I/O devices, NICs, and network aspects of the system such as TCP/IP, Netbeui, Novell, etc. Further, management application 32 manages the software applications running on computer system 110 by determining what is currently running on the system, what is currently installed, the state of installation, which applications can be terminated, performing application metering, managing application life cycle, process management, user management, etc.

[0031] In one embodiment of the invention, developer 30 uses a CIM workshop 152 written in JAVA for viewing, changing, adding and deleting CIM classes and instances. CIM workshop 152 provides a graphical user interface for the developer. For example, developer 30 may view and select namespaces, may add namespaces, add properties, qualifiers and methods to new classes, view and create instances, and view and modify instance values. Developer 30 may also use CIM workshop 152 to browse a class inheritance tree and change the root of an object tree for a namespace.

[0032] MOF compiler 154 pares files created in the Managed Object Format (MOF), converts files to JAVA class and stores the extracted classes and instances in repository 130. The MOF language is a syntax for defining CIM classes and instances and is described in the CIM specification. Although classes and instances can also be added through client API 156 using JAVA, MOF compiler 154 eliminates the need to write such code. Compiler 154 provides developers and administrators with a simple and fast technique for modifying repository 130.

[0033] In one embodiment, client API 156 is a public API that JAVA applications use to request operations from object manager 20. Client API 156 is used by management application 32 to transfer data to and from object manager 20. Client API 156 includes a variety of classes, instances and methods useful for communicating with object manager 20 using any suitable transport mechanism.

[0034] Preferably, Client API 156 is an application programming interface used by management application 32 to communicate with object manager 20 using Remote Method Invocation (RMI) protocol 158 or XML over an HTTP protocol 160 according to the techniques described in U.S. patent application No. ________________ (Atty Docket SUN1P366) referenced above. Other suitable protocols may also be used such as COM from Microsoft Corporation. Client API 156 may communicate directly with object manager 20 using RMI or may communicate using the XML/HTTP protocol using web server 140. Alternatively, client API 156 can communicate using the XML/HTTP protocol 160 directly should object manager 20 support the HTTP format.

[0035] Connections 170a-170d are any suitable local or network connection between computer system 110 and application computer 150. By way of example, these connections occur over an internet, an intranet, an extranet, within a workgroup, or other.

[0036] FIG. 2B illustrates further detail in which CIM object manager 20 communicates with any of a variety of remote CIM repositories using a repository API 180. In this embodiment, each of repositories 190, 192 and 194 are located remotely from object manager 20 and computer system 110. In this example each repository is located on a different computer although it is conceivable that all may be located on a single computer, or that a repository is local to computer system 110.

[0037] Repository 190 is a database implemented using a flat file technique or object serialization in JAVA; it communicates with Repository API 180 over a network connection 132a that uses simple JAVA code protocol. Repository 192 is an object-oriented database and may be implemented using tools such as those available from Sybase, Oracle, or Informix. Repository 812 communicates with Repository API 180 over a network connection 132b using a JAVA Database Connectivity (JDBC) protocol. Repository 194 is a Lightweight Directory Access Protocol (LDAP) type of database that communicates over network connection 132c using a JAVA Naming Directory Interface (JNDI) protocol.

[0038] Repository API 180 is used by object manager 20 to store data to, or retrieve data from, the repositories. Repository API 180 includes a variety of classes, instances and methods useful for communicating with the repositories using any suitable protocol. Preferably, Repository API 180 communicates with the repositories using a JAVA language protocol, a JDBC protocol, a JNDI protocol, an LDAP protocol, an ODBC protocol, or other protocols suitable for use with a database. Implementation of such communication between object manager 20 and the repositories according to an embodiment of the invention is further described in FIGS. 4-8.

[0039] FIGS. 3A and 3B illustrate an example graphical user interface for CIM workshop 152. Preferably, a login to the workshop prompts for a host name,

namespace, user name and password. By default, workshop 152 connects to the object manager on the local host in the default namespace. FIG. 3A illustrates CIM classes that represent objects in the selected namespace on the selected host. Listed in panel 210 are the objects of the selected namespace. On the right-hand panel are shown the properties 212 for the selected object (in this case the object "Solaris Package") and methods 214 (not shown). FIG. 3B shows all instances of a selected object. Instances are shown in the left-hand panel, and in this example instance 252 is shown. The right-hand panel shows all properties 254 associated with the selected instance and its associated methods 256 (not shown).

[0040] FIG. 4 illustrates an interface definition 300 of Repository API 180 useful for implementing an embodiment of the invention. Interface 300 lists various methods that may be called by object manager 20 in the course of database operations with repositories 190-194. Advantageously, this interface may be implemented using a variety of classes having protocol-specific methods thus allowing a transport neutral object manager to be written. Interface 300 includes an interface name 302 which in this example is "CIM Repository API." Included are a variety of methods defined for the interface. Each method has a method name 304, a return value 306 and parameters 308. By way of example, shown is one method "Add CIM Element" 310 having a return value of "void" and accepting parameters "element" and "namespace." A large number of other methods may be defined for interface 300

[0041] By way of example, these methods include the following. The Create Namespace method creates a CIM namespace, a directory containing classes and instances. (When a management application connects to object manager 20 it specifies a namespace. All subsequent operations occur within that namespace on the object manager host.) The method Delete Class deletes the specified class. The method Delete Instance deletes the specified instance. The method Delete Qualifier deletes the specified qualifier. The method Enumerate Classes retrieves the specified classes from a repository. The method Enumerate Namespace gets a list of namespaces. The method Enumerate Instances gets of list of instances for the specified class. The method Enumerate Qualifier Types get a list of qualifier types for the specified class. The method Get Class gets the CIM class for the specified CIM object path. The Get Instance method gets the CIM instance for the specified CIM object path.

[0042] The method Get Qualifier Type gets the qualifier type for the specified CIM object path. The method Set Instance invokes a repository to add or update the specified CIM instance to the specified namespace. Other methods may also be included within interface 300 such as Add Aliased Class Name, Add Aliased Instance Name, Get Aliased Class Name, Get Aliased Instance Name, etc.

[0043] Once interface 300 has been defined it is possible to then code protocol-specific methods to implement each of the methods defined in interface 300. In this fashion, any number of protocol-specific classes are provided each having an implementation for a specific protocol such as JDBC or LDAP. Though the use of these protocol-specific classes, object manager 20 is able to communicate with any CIM repository using any suitable protocol in a transparent fashion.

[0044] FIG. 5 illustrates an implementation definition 400 for the interface of FIG. 4. Specifically, implementation 400 implements class "CIM Repository API" using methods specific to the LDAP protocol. Such protocol-specific methods allow object manager 20 to communicate via Repository API 180 to repository 194 using the LDAP protocol. Implementation 400 includes a class name 402 "CIM Repository LDAP." Also included is constructor definition code 404 that constructs an instance of class 402 that is specific to the LDAP protocol. Use of a constructor definition to create an instance is well known to those of skill in the art.

[0045] Also included in implementation 400 are the specific implementations of the methods defined upon interface 300. For each method implemented there is a method name 406, a return value 408, parameters 410 and implementation code 412. Implementation code 412 is preferably JAVA code that implements the particular method using any constructs necessary that are specific to the RMI protocol. Those of skill in the art will appreciate how to implement JAVA code for a particular purpose that must adhere to a specific protocol.

[0046] Preferable, all of the methods defined upon interface 300 are implemented in implementation 400. Shown by way of example is the method Add CIM Element 416 which has a return value of "void" and accepts the parameters element and namespace. Not shown for simplicity is the actual LDAP-specific JAVA code that implements the method Add CIM Element. The other methods defined in interface 300 are also listed in implementation 400 along with their LDAP-specific code.

[0047] FIG. 6 illustrates an implementation definition 500 for the interface of FIG. 4. Specifically, implementation 500 implements class "CIM Repository API" using methods specific to the JDBC protocol. Such protocol-specific methods allow object manager 20 to communicate via Repository API 180 to repository 192 using the JDBC protocol. Implementation 500 includes a class name 502 "CIM Repository JDBC." Also included is constructor definition code 504 that constructs an instance of class 502 that is specific to the JDBC protocol. Use of a constructor definition to create an instance is well known to those of skill in the art.

[0048] Also included in implementation 500 are the specific implementations of the methods defined upon interface 300. For each method implemented there is a method name 506, a return value 508, parameters 510 and implementation code 512. Implementation code

512 is preferably JAVA code that implements the particular method using any constructs necessary that are specific to the JDBC protocol. Those of skill in the art will appreciate how to implement JAVA code for a particular purpose that must adhere to a specific protocol.

[0049] Preferable, all of the methods defined upon interface 300 are implemented in implementation 500. Shown by way of example is the method Add CIM Element 516 which has a return value of "void" and accepts the parameters element and namespace. Not shown for simplicity is the actual JDBC-specific JAVA code that implements the method Add CIM Element. The other methods defined in interface 300 are also listed in implementation 500 along with their JDBC-specific code.

[0050] FIG. 7 is a JAVA factory class 600. Factory 600 is used for determining which protocol is desired by object manager 20 and directing the creation of a protocol-specific object to be returned to the object manager.

[0051] Factory 600 includes a class name 602 "CIM Repository Factory" and any number of defined methods. For each method there is a method name 604, a return value 606, parameters 608 and an implementation 610. In particular, the method Get Repository API accepts the parameters protocol, namespace and version, and returns an instance of interface 300 which is a protocol-specific instance of either implementation 400 or implementation 500. Of course, other protocol-specific objects may be returned if other implementations are defined. The implementation code 610 for method 612 may be any suitable JAVA code that checks the protocol parameter to see which protocol is desired and then directs either implementation 400 or 500 to construct a new instance of itself. By way of example, a series of case statements may be used. Other methods may also be defined and implemented within factory 600.

OBJECT MANAGER EXECUTION

[0052] FIG. 8 is a flowchart illustrating invocation of a method by the object manager to perform a database operation such as storing or retrieving an object. Once management application 32 has been created by developer 30 and the classes and methods of FIGS. 4-7 have been defined, the object manager may perform operations on one of the repositories using any desired protocol. FIG. 8 illustrates a single method call according to one embodiment of the invention.

[0053] In step 702 management application 32 creates a connection from application computer 150 to computer system 110. Preferably, application 32 invokes a method within Client API 156 which creates an instance of application 32 within object manager 20. Application 32 passes to the method a host name, a namespace, a user name, a password, and the protocol by which it is desired to communicate with host computer system 110. Any suitable network protocol may be identified such as RMI, XML/HTTP or DCOM.

[0054] In step 704 object manager 20 receives a method call from application 32 that requires a database operation. The method call is preferably performed using the technique described in U.S. patent application No. 09/333,878. In response to this method call, object manager 20 identifies a repository and protocol and makes a call to Repository API 180.

[0055] In step 706 factory 600 of Repository API 180 checks the protocol desired by object manager 20 using its method Get Repository API. This method returns a protocol-specific object which is an instance of either the class defined in implementation 400 or the class defined in implementation 500. For example, in step 710 if the protocol parameter is LDAP, then in step 714 the constructor definition 404 of implementation 400 executes and results in an LDAP-specific object having LDAP-specific methods being returned to the object manager. On the other hand, if the protocol parameter is JDBC, then in step 722 the constructor definition 504 of implementation 500 executes and produces a JDBC-specific object which is returned to the object manager. As shown in steps 726 and 730, a desire for use of simple JAVA protocol results in a JAVA-specific object being returned. Other protocols are also supported. In step 738 object manager 20 invokes a desired database method upon the protocol-specific object recently returned. Because the methods of this object are specific to the protocol desired by object manager 20, communication between Repository API 180 and the target repository occurs using the desired protocol in a fashion transparent to application 32 and to object manager 20.

[0056] Once the target CIM repository has processed the method (which may be a request for an object, a request to add an object, etc.), then in step 742 the result is returned from the repository to object manager 20 via Repository API 180 using the desired protocol. In this fashion, a technique has been described that allows an object manager to be written independent of the protocol by which it is desired to communicate with a target CIM repository.

COMPUTER SYSTEM EMBODIMENT

[0057] FIGS. 9 and 10 illustrate a computer system 900 suitable for implementing embodiments of the present invention. FIG. 9 shows one possible physical form of the computer system. Of course, the computer system may have many physical forms ranging from an integrated circuit, a printed circuit board and a small handheld device up to a huge super computer. Computer system 900 includes a monitor 902, a display 904, a housing 906, a disk drive 908, a keyboard 910 and a mouse 912. Disk 914 is a computer-readable medium used to transfer data to and from computer system 900.

[0058] FIG. 10 is an example of a block diagram for computer system 900. Attached to system bus 920 are

a wide variety of subsystems. Processor(s) 922 (also referred to as central processing units, or CPUs) are coupled to storage devices including memory 924. Memory 924 includes random access memory (RAM) and read-only memory (ROM). As is well known in the art, ROM acts to transfer data and instructions uni-directionally to the CPU and RAM is used typically to transfer data and instructions in a bi-directional manner. Both of these types of memories may include any suitable of the computer-readable media described below. A fixed disk 926 is also coupled bi-directionally to CPU 922; it provides additional data storage capacity and may also include any of the computer-readable media described below. Fixed disk 926 may be used to store programs, data and the like and is typically a secondary storage medium (such as a hard disk) that is slower than primary storage. It will be appreciated that the information retained within fixed disk 926, may, in appropriate cases, be incorporated in standard fashion as virtual memory in memory 924. Removable disk 914 may take the form of any of the computer-readable media described below.

[0059] CPU 922 is also coupled to a variety of input/output devices such as display 904, keyboard 910, mouse 912 and speakers 930. In general, an input/output device may be any of: video displays, track balls, mice, keyboards, microphones, touch-sensitive displays, transducer card readers, magnetic or paper tape readers, tablets, styluses, voice or handwriting recognizers, biometrics readers, or other computers. CPU 922 optionally may be coupled to another computer or telecommunications network using network interface 940. With such a network interface, it is contemplated that the CPU might receive information from the network, or might output information to the network in the course of performing the above-described method steps. Furthermore, method embodiments of the present invention may execute solely upon CPU 922 or may execute over a network such as the Internet in conjunction with a remote CPU that shares a portion of the processing.

[0060] In addition, embodiments of the present invention further relate to computer storage products with a computer-readable medium that have computer code thereon for performing various computer-implemented operations. The media and computer code may be those specially designed and constructed for the purposes of the present invention, or they may be of the kind well known and available to those having skill in the computer software arts. Examples of computer-readable media include, but are not limited to: magnetic media such as hard disks, floppy disks, and magnetic tape; optical media such as CD-ROMs and holographic devices; magneto-optical media such as floptical disks; and hardware devices that are specially configured to store and execute program code, such as application-specific integrated circuits (ASICs), programmable logic devices (PLDs) and ROM and RAM devices. Examples of computer code include machine code, such as produced by a compiler, and files containing higher level code that are executed by a computer using an interpreter.

[0061] Although the foregoing invention has been described in some detail for purposes of clarity of understanding, it will be apparent that certain changes and modifications may be practiced within the scope of the appended claims. For instance, the application computer and the computer system to be managed may be the same computer, or may be separated by a great distance. Also, the various CIM repositories may be located along with the computer system, may each be remotely located on a separate computer, or may be remotely located on a single computer. The use of a web server may not be required should the CIM object manager support the HTTP format. Other types of classes and methods may be used while not departing from the spirit of the invention. Therefore, the described embodiments should be taken as illustrative and not restrictive, and the invention should not be limited to the details given herein but should be defined by the following claims and their full scope of equivalents.

Claims

1. A method for communication between a Common Information Model (CIM) object manager of a host computer and a CIM repository, said method comprising:

creating a connection between said object manager and said CIM repository;

passing a protocol indicator from said object manager to a repository application programming (API), said protocol indicator identifying a protocol by which said CIM object manager desires to communicate with said CIM repository;

creating a protocol-specific object having methods implemented using said protocol; and

returning said protocol-specific object to said CIM object manager, whereby said CIM object manager may communicate with said CIM repository using said protocol.

2. The method of claim 1 further comprising:

invoking a method defined upon said protocol-specific object;

transmitting said method using said protocol over said connection to said CIM repository; and

returning a result to said CIM object manager over said connection using said protocol.

3. The method of claim 1 wherein said protocol is LDAP, JDBC, or JAVA.

4. The method of claim 1 wherein said CIM repository is resident on said host computer.

5. The method of claim 1 wherein said CIM repository is resident on a separate computer.

6. The method of claim 1 wherein said creating a protocol-specific object includes

calling a JAVA factory class.

7. A computer system for interacting with a CIM repository database, said system comprising:

a CIM object manager including a protocol indicator and program code for interacting with said CIM repository; and

a repository application programming interface (repository API) including

a factory class arranged to receive said protocol indicator from said object manager and produce a protocol-specific object,

a first class having methods defined thereon implemented in a first protocol, and

a second class having methods defined thereon implemented in a second protocol, whereby said protocol-specific object may be returned to said object manager for use in interacting with said CIM repository.

8. The system of claim 7 wherein said CIM object manager is arranged to receive a method call from a management application using the protocol identified by said protocol indicator.

9. The system of claim 7 wherein said CIM repository is resident on said computer system.

10. The system of claim 7 wherein said computer system and said CIM repository are connected over a network connection implemented in the protocol identified by said protocol indicator.

11. The system of claim 7 wherein the protocol identified by said protocol indicator is LDAP, JDBC or JAVA.

12. The system of claim 7 further comprising:

a plurality of CIM repositories, each repository arranged to communicate with said CIM object manager using a different protocol.

13. The system of claim 12 wherein each repository is resident on a different computer.

14. A computer-readable medium comprising computer code for communication between a Common Information Model (CIM) object manager of a host computer and a CIM repository, said computer code of said computer-readable medium effecting the following:

creating a connection between said object manager and said CIM repository;

passing a protocol indicator from said object manager to a repository application programming (API), said protocol indicator identifying a protocol by which said CIM object manager desires to communicate with said CIM repository;

creating a protocol-specific object having methods implemented using said protocol; and

returning said protocol-specific object to said CIM object manager, whereby said CIM object manager may communicate with said CIM repository using said protocol.

15. The computer-readable medium of claim 14 further comprising computer code for effecting the following:

invoking a method defined upon said protocol-specific object;

transmitting said method using said protocol over said connection to said CIM repository; and

returning a result to said CIM object manager over said connection using said protocol.

16. The computer-readable medium of claim 14 wherein said protocol is LDAP, JDBC or JAVA.

17. The computer-readable medium of claim 14 wherein said creating a protocol-specific object includes

calling a JAVA factory class.

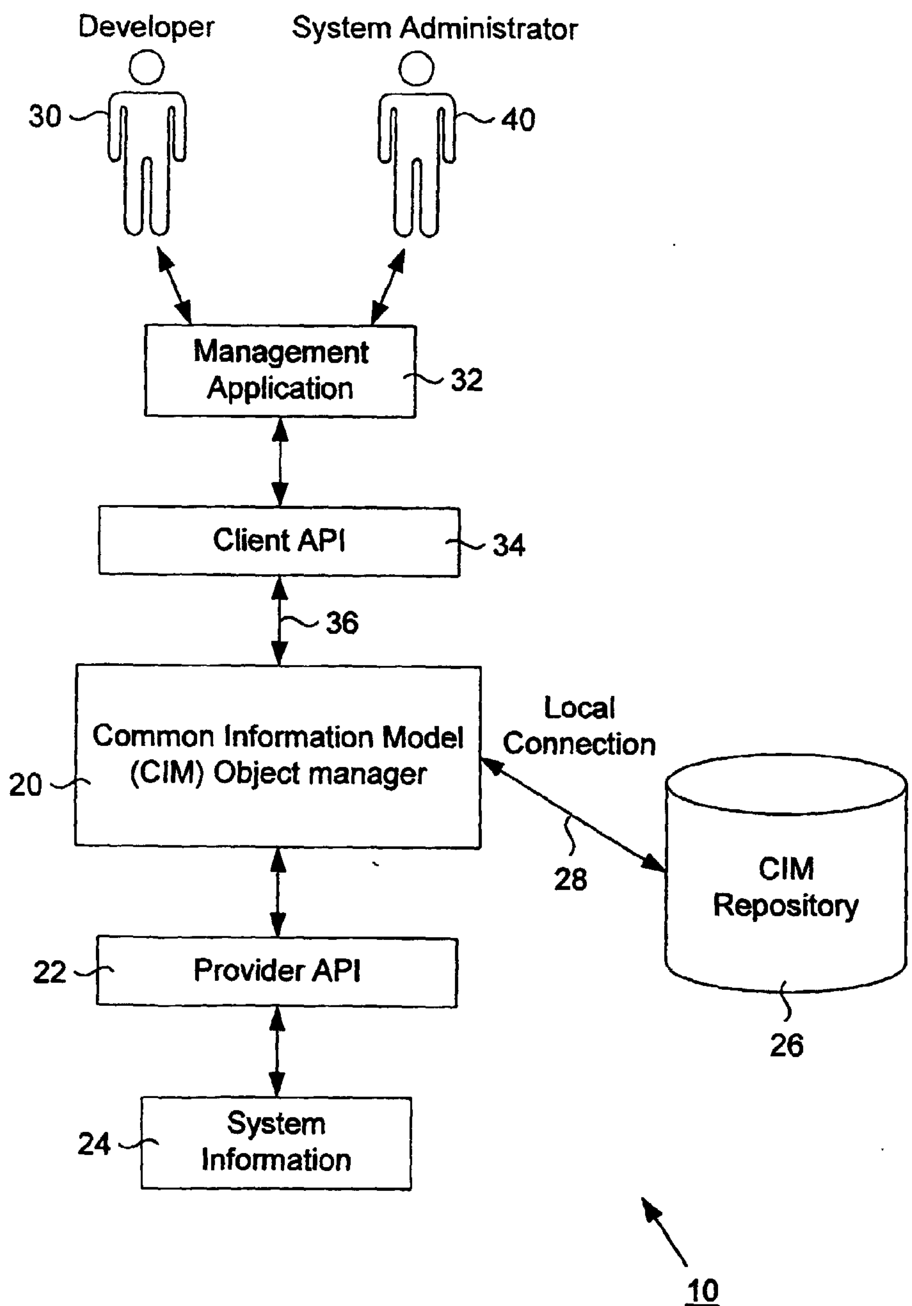


FIG. 1
(Prior Art)

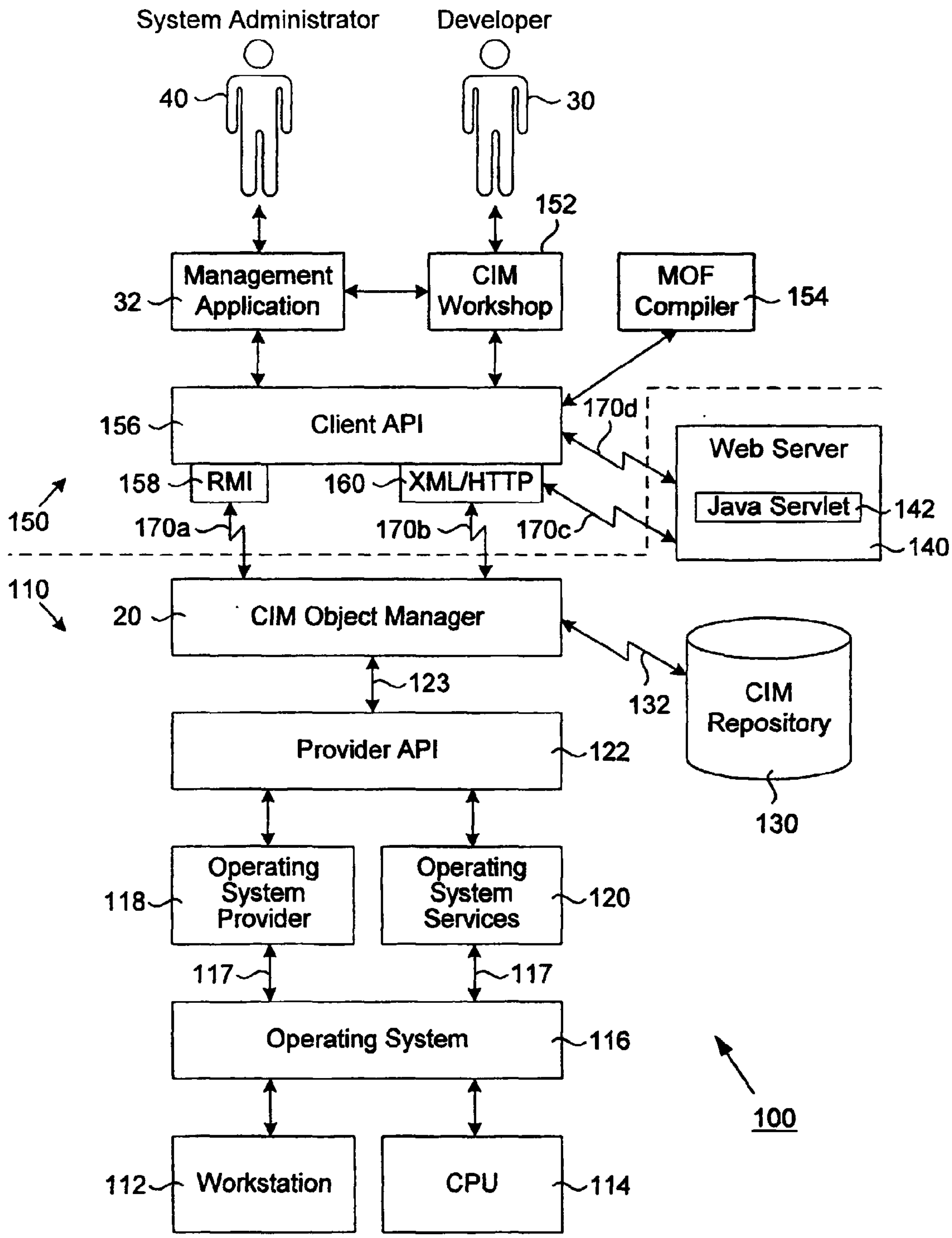

System Administrator
Developer
40
30
152
Management Application
CIM Workshop
MOF Compiler
32
154
Client API
156
170d
Web Server
Java Servlet
142
140
158
RMI
160
XML/HTTP
150
170a
170b
170c
110
20
CIM Object Manager
123
132
CIM Repository
Provider API
122
130
Operating System Provider
Operating System Services
118
120
117
117
Operating System
116
100
Workstation
CPU
112
114

FIG. 2A

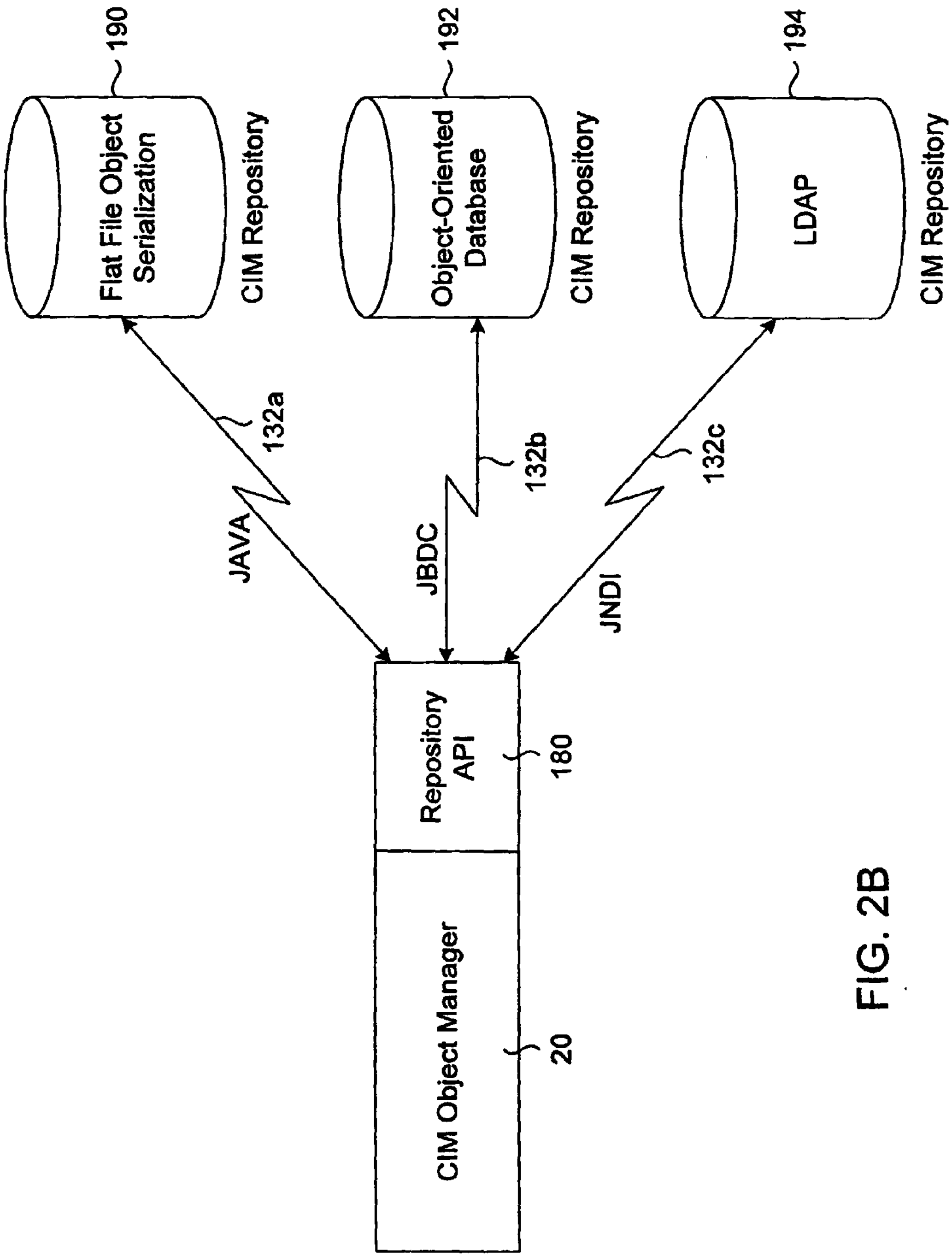
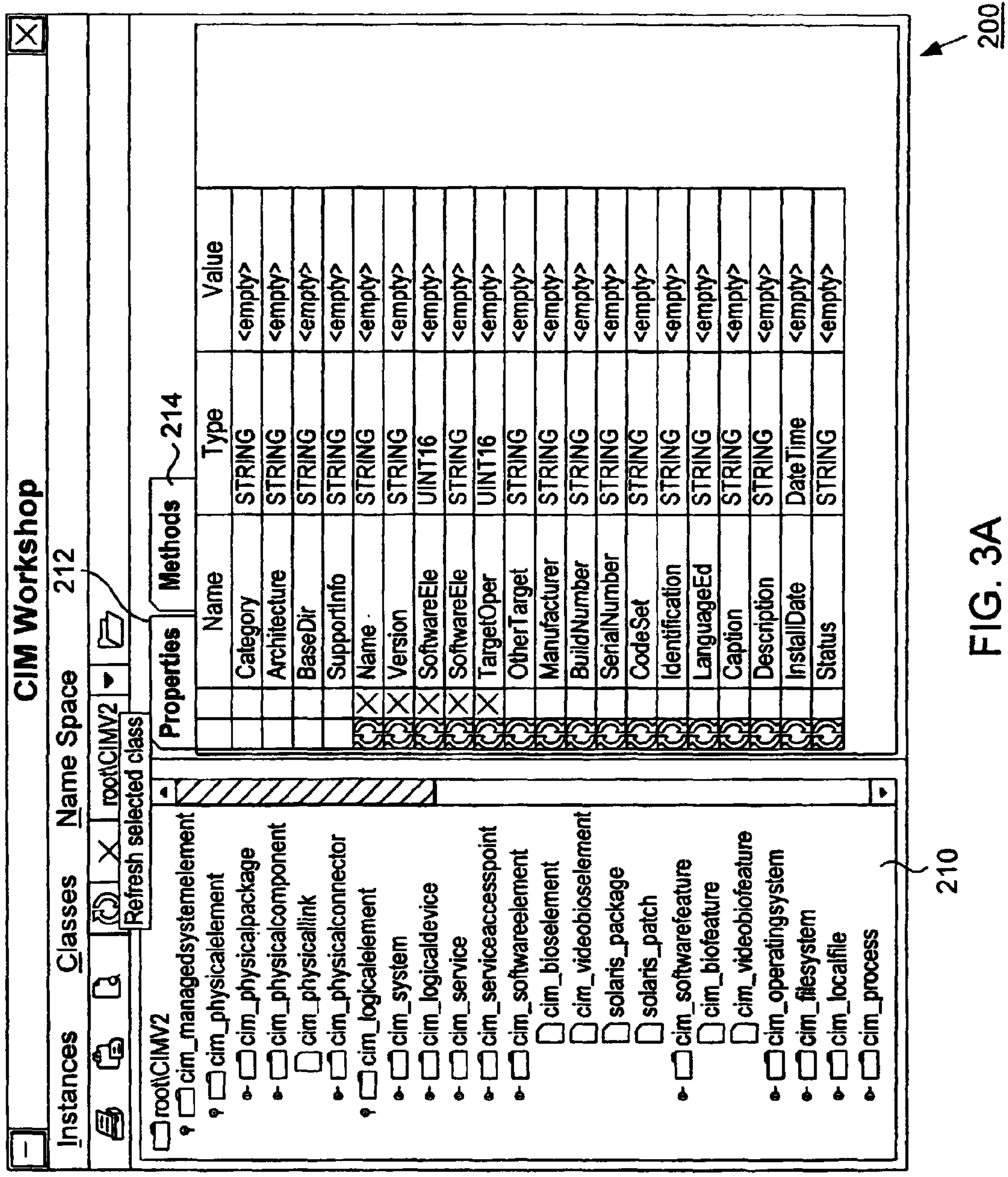
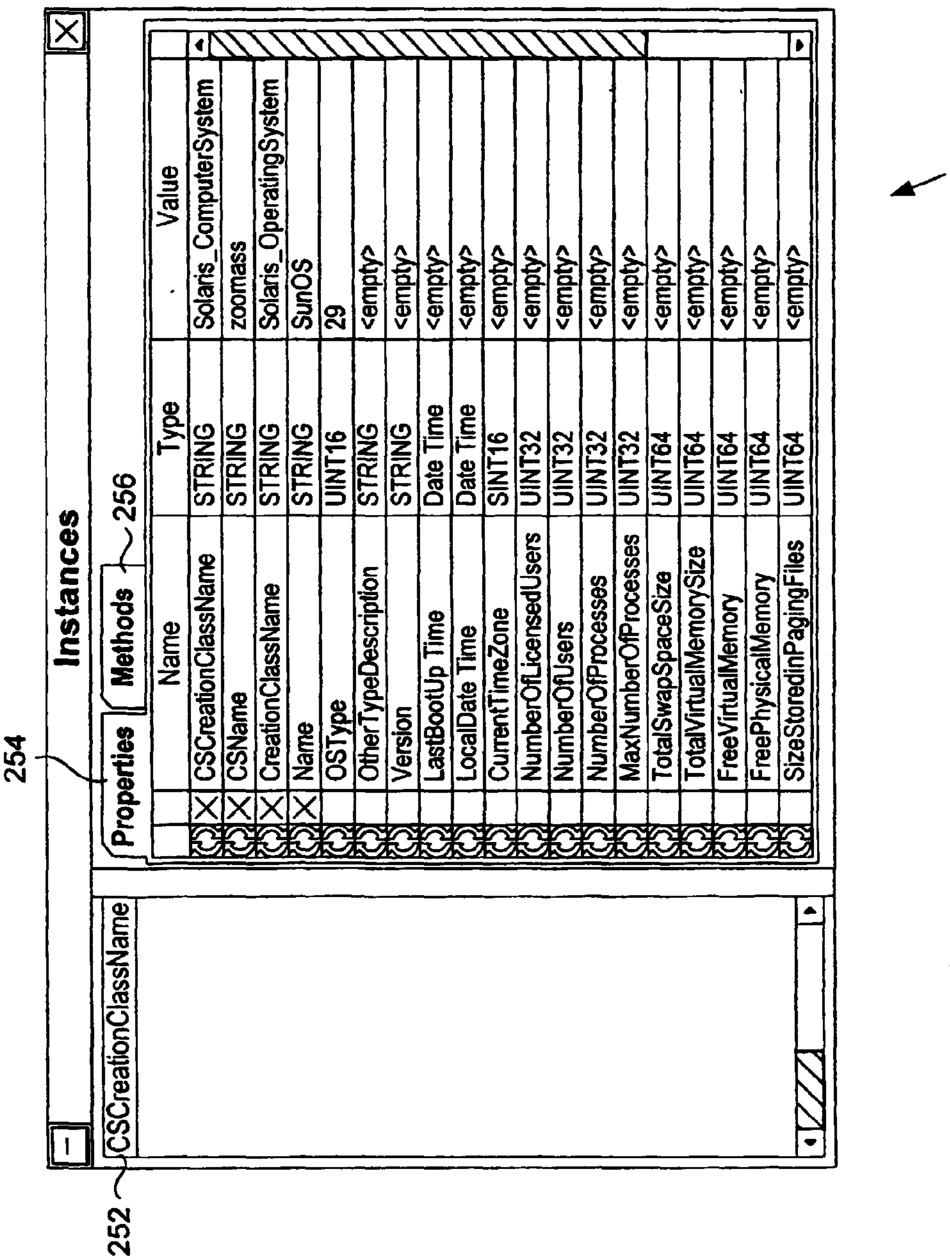

Flat File Object Serialization
190
CIM Repository
Object-Oriented Database
192
CIM Repository
LDAP
194
CIM Repository
132a
JAVA
132b
JBDC
132c
JNDI
Repository API
180
CIM Object Manager
20

FIG. 2B



CIM Workshop
Instances
Classes
Name Space
root\CIMV2
Refresh selected class
212
214
Properties
Methods
root\CIMV2
cim_managedsystemelement
cim_physicalelement
cim_physicalpackage
cim_physicalcomponent
cim_physicallink
cim_physicalconnector
cim_logicalelement
cim_system
cim_logicaldevice
cim_service
cim_serviceaccesspoint
cim_softwareelement
cim_bioselement
cim_videobioselement
solaris_package
solaris_patch
cim_softwarefeature
cim_biofeature
cim_videobiofeature
cim_operatingsystem
cim_filesystem
cim_localfile
cim_process
210
Name
Type
Value
Category STRING <empty>
Architecture STRING <empty>
BaseDir STRING <empty>
SupportInfo STRING <empty>
Name STRING <empty>
Version STRING <empty>
SoftwareEle UINT16 <empty>
SoftwareEle STRING <empty>
TargetOper UINT16 <empty>
OtherTarget STRING <empty>
Manufacturer STRING <empty>
BuildNumber STRING <empty>
SerialNumber STRING <empty>
CodeSet STRING <empty>
Identification STRING <empty>
LanguageEd STRING <empty>
Caption STRING <empty>
Description STRING <empty>
InstallDate DateTime <empty>
Status STRING <empty>
200

FIG. 3A


Instances
CSCreationClassName
Properties
Methods
Name
Type
Value
CSCreationClassName
STRING
Solaris_ComputerSystem
CSName
STRING
zoomass
CreationClassName
STRING
Solaris_OperatingSystem
Name
STRING
SunOS
OSType
UINT16
29
OtherTypeDescription
STRING
<empty>
Version
STRING
<empty>
LastBootUp Time
Date Time
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LocalDate Time
Date Time
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CurrentTimeZone
SINT16
<empty>
NumberOfLicensedUsers
UINT32
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NumberOfUsers
UINT32
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NumberOfProcesses
UINT32
<empty>
MaxNumberOfProcesses
UINT32
<empty>
TotalSwapSpaceSize
UINT64
<empty>
TotalVirtualMemorySize
UINT64
<empty>
FreeVirtualMemory
UINT64
<empty>
FreePhysicalMemory
UINT64
<empty>
SizeStoredinPagingFiles
UINT64
<empty>
252
254
256
250

FIG. 3B

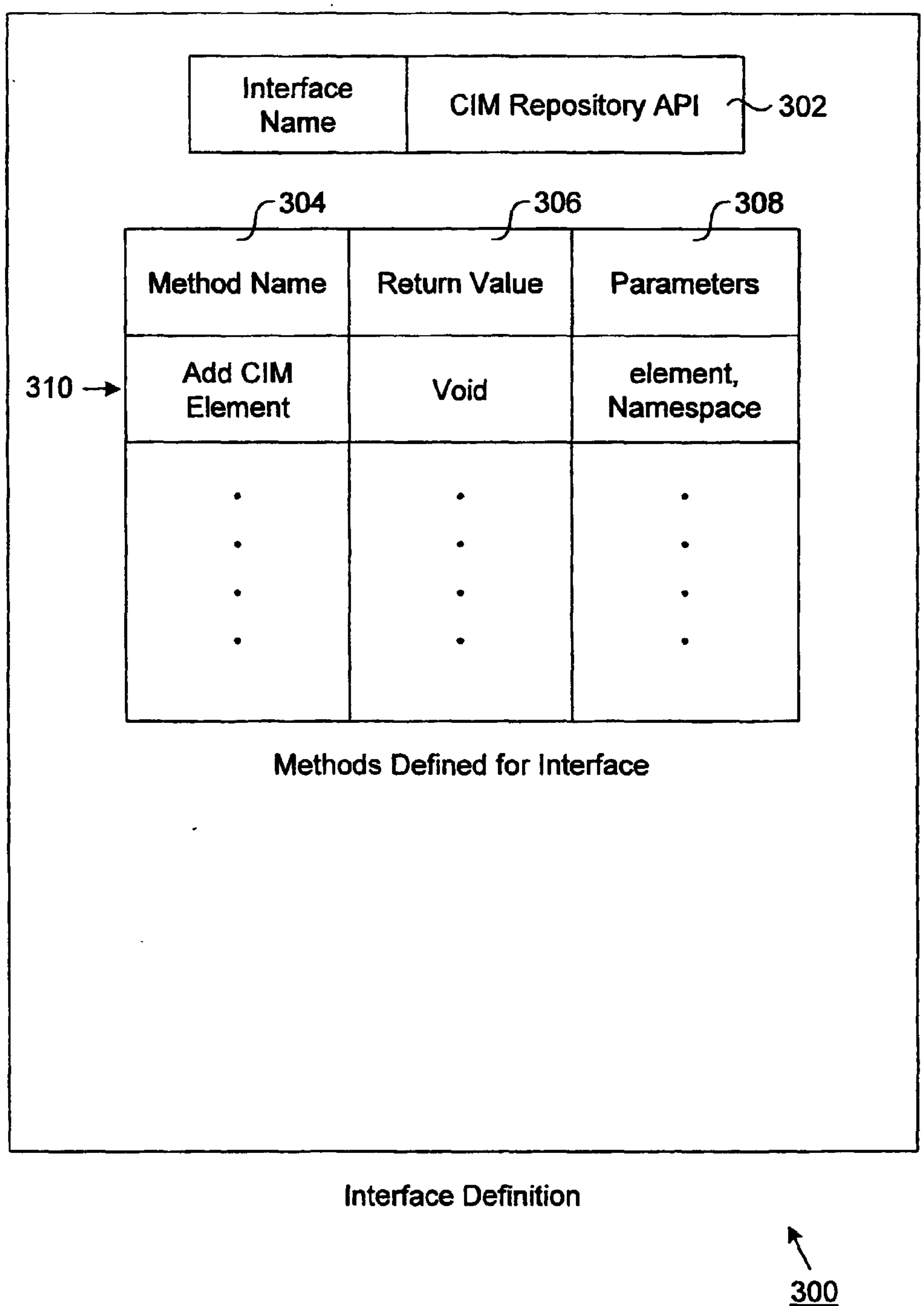
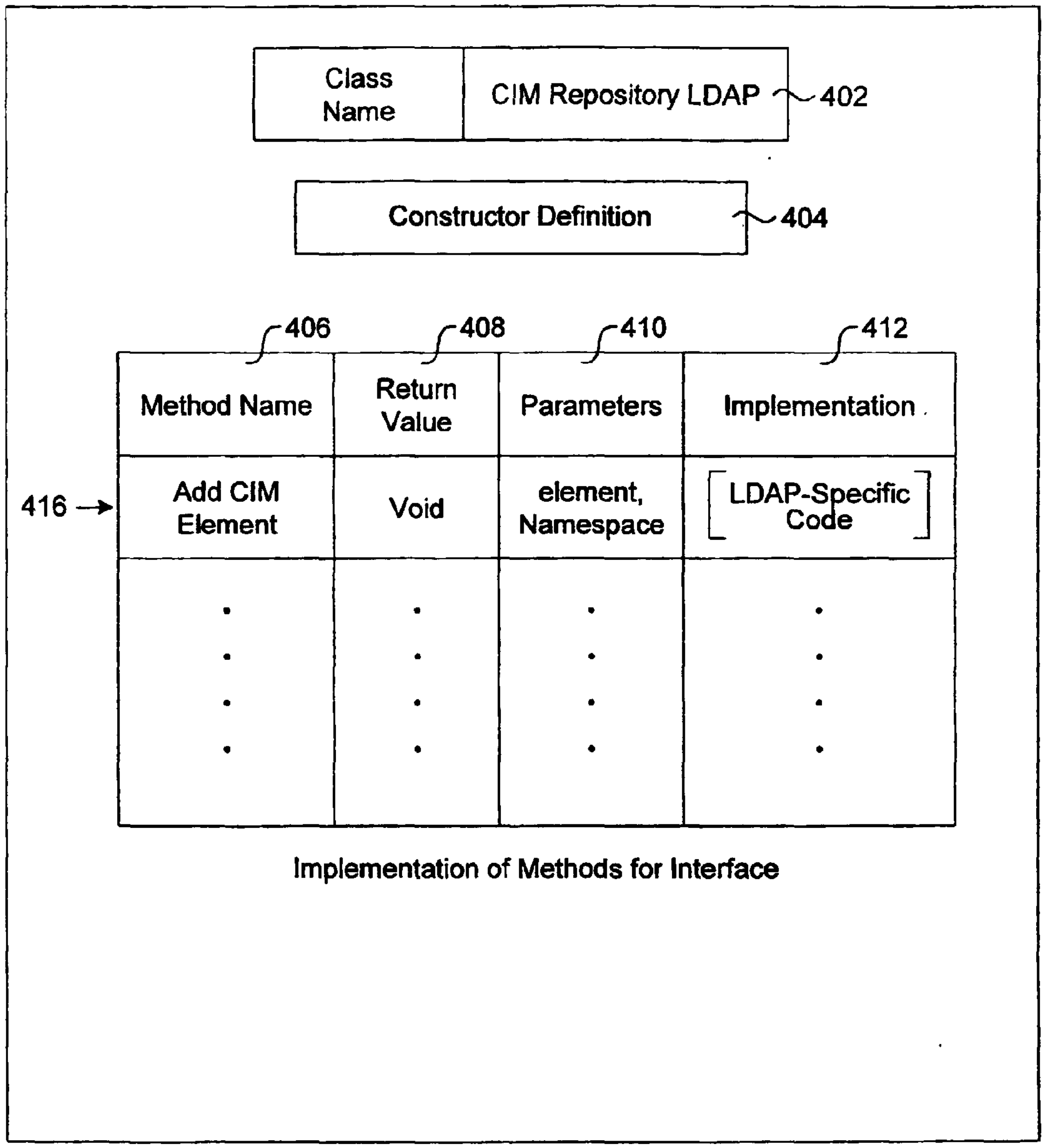

Interface Name
CIM Repository API
302
304
306
308
Method Name
Return Value
Parameters
310
Add CIM Element
Void
element, Namespace
Methods Defined for Interface
Interface Definition
300

FIG. 4



Implementation Definition for Interface

400

FIG. 5

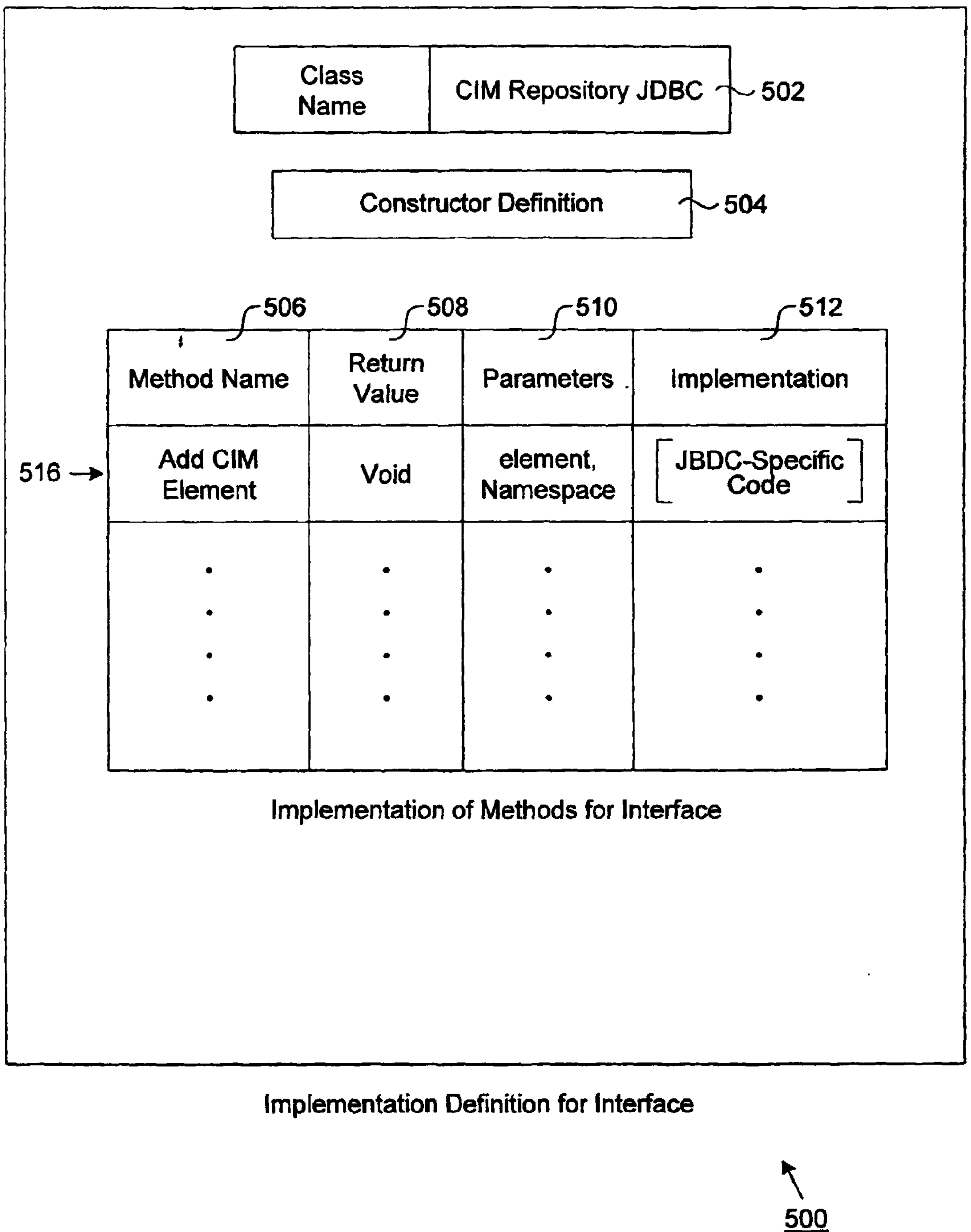

Class Name
CIM Repository JDBC
502
Constructor Definition
504
506
508
510
512
Method Name
Return Value
Parameters
Implementation
516
Add CIM Element
Void
element, Namespace
JBDC-Specific Code
Implementation of Methods for Interface
Implementation Definition for Interface
500

FIG. 6

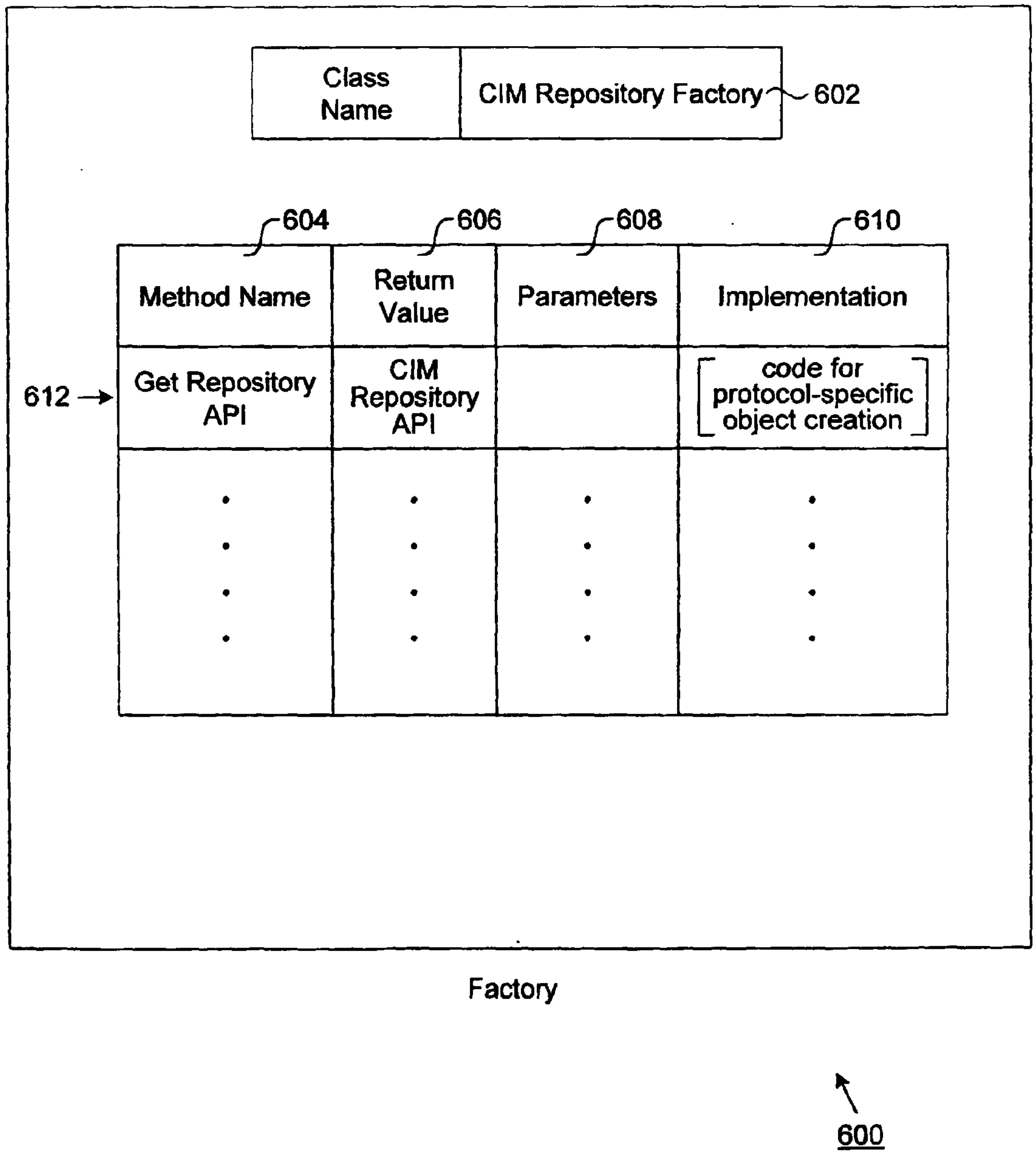

Class Name
CIM Repository Factory
602
604
606
608
610
Method Name
Return Value
Parameters
Implementation
612
Get Repository API
CIM Repository API
code for protocol-specific object creation
Factory
600

FIG. 7

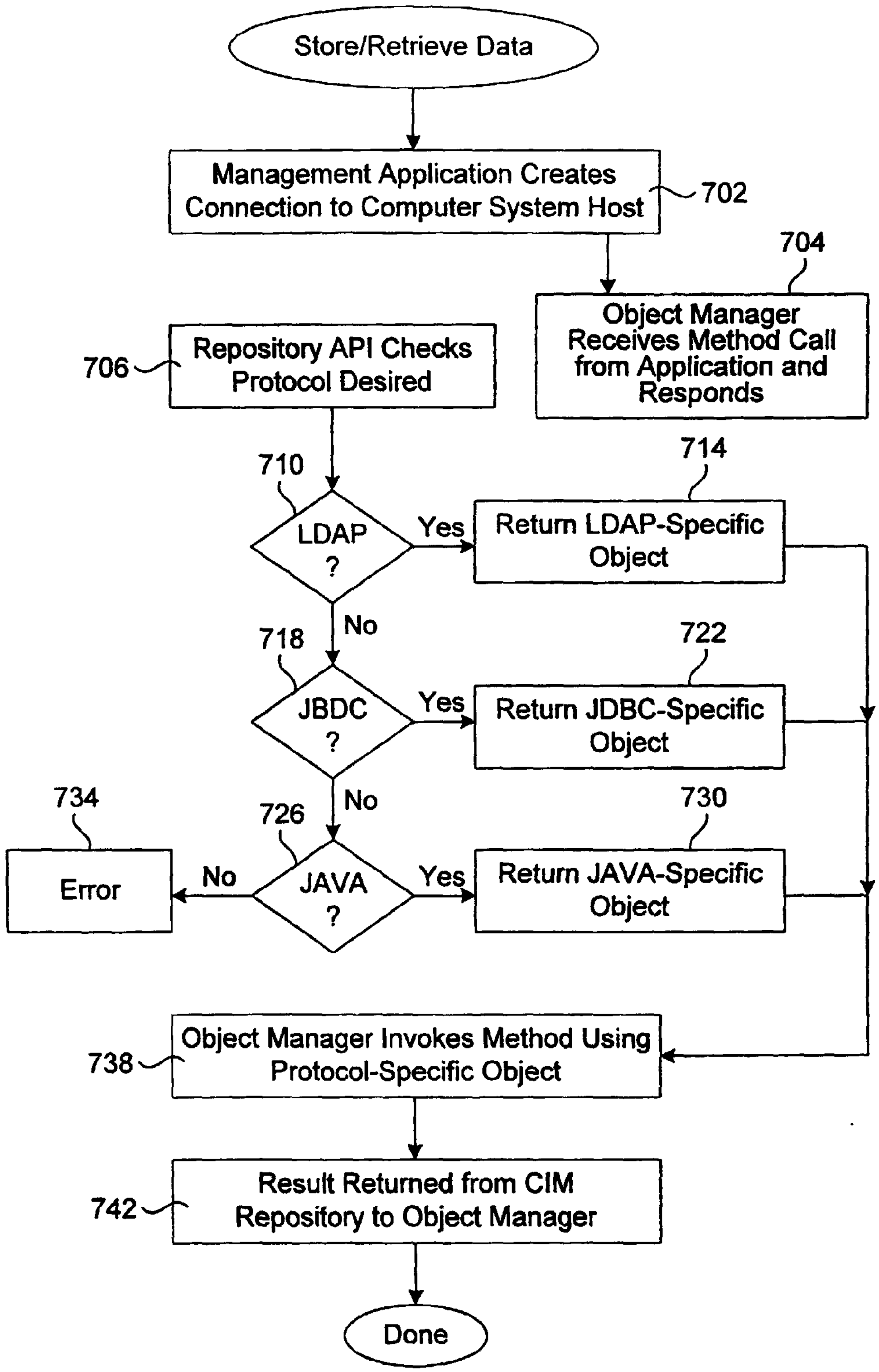

Store/Retrieve Data
Management Application Creates Connection to Computer System Host
702
704
Object Manager Receives Method Call from Application and Responds
706
Repository API Checks Protocol Desired
710
LDAP ?
Yes
714
Return LDAP-Specific Object
No
718
JBDC ?
Yes
722
Return JDBC-Specific Object
No
734
726
Error
No
JAVA ?
Yes
730
Return JAVA-Specific Object
738
Object Manager Invokes Method Using Protocol-Specific Object
742
Result Returned from CIM Repository to Object Manager
Done

FIG. 8

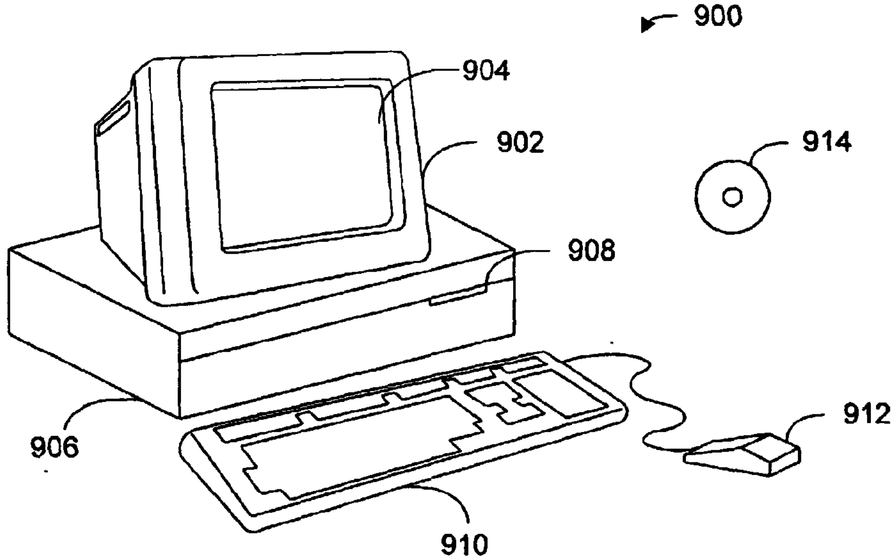

900
904
902
914
908
906
912
910

FIG. 9

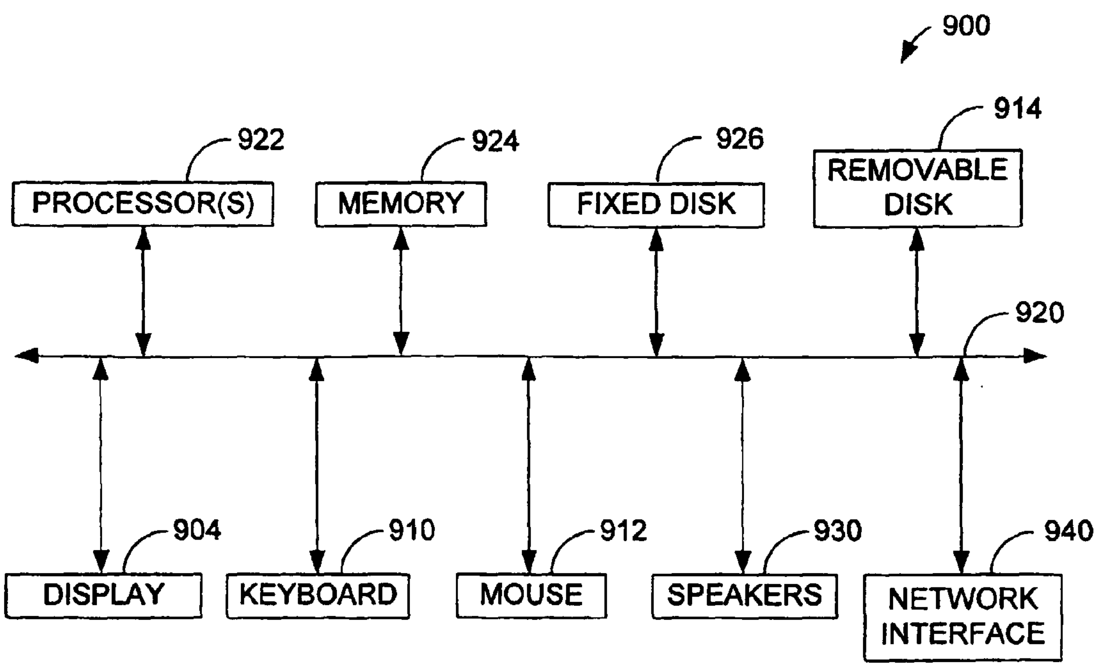

900
922
PROCESSOR(S)
924
MEMORY
926
FIXED DISK
914
REMOVABLE DISK
920
904
DISPLAY
910
KEYBOARD
912
MOUSE
930
SPEAKERS
940
NETWORK INTERFACE

FIG. 10

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(71) Applicant: **INTERUNIVERSITAIR MICRO-ELEKTRONICA CENTRUM VZW**
3001 Heverlee (BE)

(72) Inventor: **Vandenameele, Patrick**
3000 Leuven (BE)

(74) Representative: **Van Malderen, Joelle et al**
Office Van Malderen,
Place Reine Fabiola 6/1
1083 Bruxelles (BE)

(54) **A method and apparatus for multi-user transmission**

(57) The present invention is related to a method of transmitting data signals (50) from at least two transmitting terminals (20) with each at least one transmitting means (60) to at least one receiving terminal (40) with a spatial diversity receiving means (80) comprising the steps:

- transmitting from said transmitting terminals (20) transformed data signals (70), being transformed versions of said data signals; receiving on said spatial diversity means (80) received data signals being at least function of at least two of said transformed data signals (70);
- subband processing (90) of at least two of said received data signals in said receiving terminal (40); and
- determining estimates of said data signals (120) from subband processed received data signals (140) in said receiving terminal.

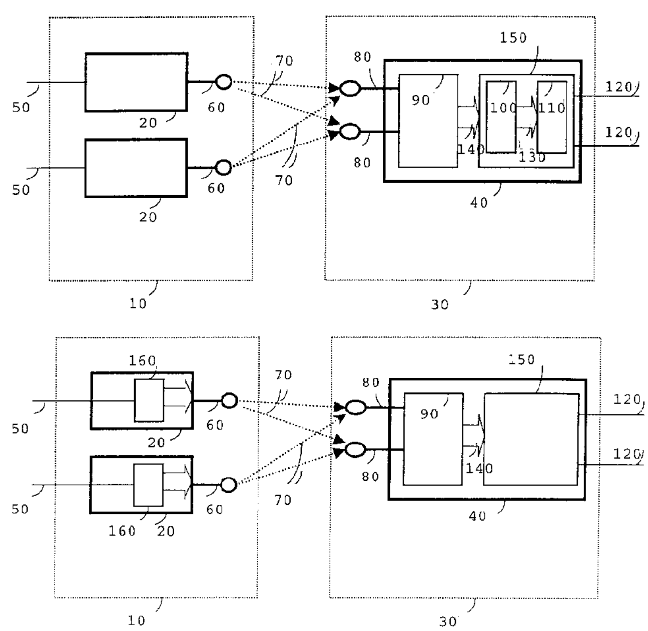

FIG. 1

EP 1 075 093 A1

Description

Field of the invention

[0001] The present invention relates to an apparatus and methods for high-speed multi-user wireless communication.

State of the art

[0002] The idea to use multicarrier methods as a modulation technique is known in the art [Chang, R.W. "Synthesis of band-limited orthogonal signals for multichannel data transmission," Bell syst. Tech. J., vol.45, pp. 1775-1796, Dec. 1966], [Saltzberg, B.R. "Performance of an efficient parallel data transmission system", IEEE Trans. Comm. Technol., vol. COM-15, Dec. 1967]. Possible benefits following from multicarrier modulation have been mentioned in many articles [Meuller, T. Brueninghaus, K. and Rohling H. "Performance of Coherent OFDM-CDMA for Broadband Mobile Communications", Wireless Personal Communications 2, Kluwer Academic Publishers, 1996, pp. 295-305], [Kaiser, S. "OFDM-CDMA versus DS-CDMA: Performance Evaluation for Fading Channels", ICC '95, pp. 1722-1726]. Numerous theoretical publications have been written on said attractive modulation technique [Kalet, "The multitone Channel", IEEE Trans. Commun., vol. 37, no. 2, Feb. 1989], [Fazel, G. Fettweis, "Multi-Carrier Spread-Spectrum", Kluwer Academic Publ., 1997]. Specifically in multipath fading propagation situations, such as for example encountered in an indoor environment, multicarrier modulation is a beneficial technique. Indeed, thanks to the insertion of a guard interval containing a cyclic prefix, it enables a very efficient way of combatting ISI, being intersymbol interference. Moreover, adaptive loading techniques make it possible to considerably increase the throughput performances [L. Van der Perre, S. Thoen, P. Vandenameele, B. Gyselinckx, M. Engels. "Adaptive loading strategy for a high speed OFDM-based WLAN". In Globecom '98. Sydney, Australia, November 1998]. However, for a given carrier modulation, the bandwidth efficiency in terms of bits/sec/hertz is fixed. Given the massive growth of wireless communication and the importance of broadband services, the spectrum becomes increasingly scarce. One method to increase the capacity or the bandwidth efficiency of a wireless system, is to apply cellularization in order to reuse spectrum in different non-interfering cells. While this technique has been applied successfully in mobile telephone networks, it is -from an economic point of view-inappropriate for small- or medium-scale indoor networks as WLANs or home LANs. First of all, high operating frequencies (i.e. millimetre wave band) would be required to achieve a reasonable reuse factor [M. Chiani, D. Dardari, A. Zanella, O. Andrisano. " Service Availability of Broadband Wireless Networks for Indoor Multimedia at Millimeter Waves". In ISSSE '98. pp. 29-33, Pisa, Italy, September 1998],[T. Ithara, T. Manabe, M. Fujita, T. Matsui and Y. Sugimoto. "Research Activities on Millimeter-Wave Indoor Wireless Communications", in ICUPC '95, Tokyo, Japan, November 1995]. Secondly, cellularisation introduces an extra layer of hierarchy and complicates the protocol stack. Thirdly, cellularisation increases the installation effort. An alternative method that allows spectrum reuse and which has none of the disadvantages of cellularisation, is the application of Space Division Multiple Access (SDMA) techniques [A. Paulraj, C. Papadias. "Space-Time Processing for Wireless Communications", IEEE Signal Processing Magazine, pp. 49-83 , November 1997]. Making use of an antenna array, SDMA can separate different users communicating over the same frequency band and at the same time, by exploiting their distinct spatial signature. As such, it allows reuse within one cell of the cellularized space. SDMA has been proposed for single-carrier systems, where its benefits have been extensively proven [G. Tsoulos, M. Beach and J. MacGeehan, "Wireless personal communications for the 21st century: European technological advances in adaptive antennas", IEEE Communications Magazine, Vol. 35, No. 9, pp. 102-9, Sept 1997],[R. Roy, "An overview of smart antenna technology and its application to wireless communication systems", in IEEE International Conf. On Personal Wireless Communications, pp234-8, New York, NY, 1997],[S. Jeng, G. Xu, H. Lin and W. Vogel, "Experimental study of antenna arrays in indoor wireless applications", in Asilomar Conference on Signals, Systems and Computers, pp. 766-70 Los Alamitos, CA, 1996]. However, these single-carrier SDMA systems for high speed (e.g. 100Mbps) wireless systems demand a massive amount of processing (e.g. in the order of Gflops) [P. Vandenameele, L. Van der Perre, B. Gyselinckx, M. Engels and H. De Man, "An SDMA Algorithm for High-Speed Wireless LAN", in Globecom 98 Sydney, Australia, pp. 189-194, November 1998]. The combination of OFDM as a modulation technique with an antenna array is known in the art [G. Raleigh and J. Cioffi, "Spatio-Termporal Coding for Wireless Communication", IEEE Transaction on Communications, Vol. 46, No.3, pp. 357-366, March 1998]. However, these algorithms are limited to a single user scenario and do not enable SDMA.

Aims of the invention

[0003] The aim of the present invention is to increase the performance/cost ratio of high-speed wireless networks by providing communication methods (also denoted transmission methods, being transmitting from at least one peer and receiving on at least one other peer) and a dedicated apparatus being inherently multi-user, multi-carrier and exploiting space division multiple access principles. Said transmission methods and said apparatus enables high spec-

trum efficient communication under multipath fading conditions.

Summary of the invention

[0004] In a first aspect of the invention (Figure 1) a method of transmitting data signals from at least two transmitting terminals with each at least one transmitting means to at least one receiving terminal with a spatial diversity receiving means is disclosed. Said spatial diversity receiving means can be a plurality of at least two antenna's being spaced apart or having a different polarization. The transmitting terminals can be grouped in a composite peer. The receiving terminals can be grouped in a processing peer. The spatial diversity receiving means comprises of at least two receiving means related to each other in such a way that said receiving means provide different spatial samples of a same data signal. In a first step in said transmitting method from said transmitting terminals transmitted data signals are transmitted. This transmission can be substantially simultaneously. The spectra of said transmitted data signals can be at least partly overlapping. The transmitted data signals are transformed versions of said data signals. In a second step in said transmitting method on said spatial diversity means received data signals are received. Said received data signals are at least function of at least two of said transmitted data signals. In a third step at least two of said received data signals are subband processed in said receiving terminal. In a last step estimates of said data signals are determined from said subband processed received data signals in said receiving terminal.

[0005] Subband processing of a data signal having a data rate, comprises in principle of splitting said data signal in a plurality of data signals, with a lower data rate and modulating each of said plurality of data signals with another carrier. Said carriers are preferably orthogonal. In an embodiment said subband processing of a data signal can be realized by using serial-to-parallel convertors and using a transformation on a group of data samples of said data signal.

[0006] The transmission method exploits thus a multi-carrier approach but then in a multi-user context as at least two transmitting terminals are present. The transmission method separates different users or transmitting terminals based on the different spatial samples of the signals received on the spatial diversity means. As such the transmission method can be understood as being a Space Division Multiple Access technique but then in a multi-carrier constellation instead of single carrier. However the method is more than a straightforward concatenation of a Space Division Multiple Access technique and a multi-carrier method. Indeed such a concatenation would result in time-domain processing of the data signals while the invention exploits the inherent frequency parallelism of subband processing techniques which result in a low complexity processing of the data signals in the frequency domain, compared to plain SDMA, being a time-domain technique. This enables low complexity nonlinear processing of the data signals, which improves the performance considerably. Indeed by such nonlinear processing of the data signals the invention exploits the frequency diversity observed for the different users. It can be mentioned that the data signals to be transmitted are often at least partially independent as they are originating from different users, although the transmission method does not rely on this.

[0007] In an embodiment of this first aspect of the invention subband by subband processing is disclosed. Said subband by subband processing can also be denoted per-subband or per-carrier processing.

[0008] In another embodiment of this first aspect of the invention a method for successive interference cancellation is disclosed. Said successive interference cancellation can be but is not limited to be realized in a subband by subband processing approach.

[0009] In yet another embodiment of this first aspect of the invention a method for interference dependent state insertion is disclosed.

[0010] In still a further embodiment of this first aspect of the invention a method for exploiting coherence grouping (being subband grouping) during initialization is disclosed.

[0011] Said embodiments of said first aspect can be combined.

[0012] In a second aspect of the invention a method of transmitting data signals from at least one transmitting terminal with a spatial diversity transmitting means to at least two receiving terminals with at least one receiving means is disclosed. The transmitting terminals can be grouped in a processing peer. The receiving terminals can be grouped in a composite peer. The spatial diversity transmitting means comprises of at least two transmitting means related to each other in such a way that said transmitting means provide different spatial samples of a same data signal. In a first step of the transmitting method combined data signals are determined in said transmitting terminal. Said combined data signals are transformed versions of said data signals. In a second step of the invented transmitting method said combined data signals are inverse subband processed. In a next step said inverse subband processed combined data signals are transmitted with said spatial diversity means. The spectra of said transmitted inverse subband processed combined data signals can be at least partly overlapping. In a further step on at least one of said receiving means of at least one receiving terminal inverse subband processed received data signals are received and then estimates of said data signals are determined from said inverse subband processed received data signals. It can be mentioned that said data signals are often at least partially independent as their destinations are different users, but the invented transmission method does not rely on that. Said first and second aspect of the invention can be combined.

[0013] In a third aspect of the invention an apparatus for determining estimates of data signals from at least two at least substantially simultaneously received data signals is disclosed. Said received data signals have at least partly overlapping spectra. Said apparatus comprises at least of at least one spatial diversity receiving means, circuitry being adapted for receiving said received data signals with said spatial diversity receiving means, circuitry being adapted for subband processing at least two of said received data signals and circuitry being adapted for determining estimates of said data signals from subband processed received data signals. Said apparatus can be exploited in uplink transmission methods in the processing peer.

[0014] In an embodiment of this third aspect of the invention parallelism in the apparatus structure is disclosed. This kind of parallelism can be exploited due to the inherent frequency parallelism, typical for the invented transmission methods. Indeed said circuitry being adapted for determining estimates of said data signals from subband processed received data signals can comprises a plurality of circuits each being adapted for determining part of said estimates of said data signals based on part of the subbands of said subband processed received data signals.

[0015] In another embodiment of this third aspect of the invention further parallelism is introduced in the apparatus structure in the circuitry being adapted for receiving said received data signals.

[0016] In a fourth aspect of the invention an apparatus for transmitting inverse subband processed combined data signals comprising at least of at least one spatial diversity transmitting means, circuitry being adapted for combining data signals, circuitry being adapted for inverse subband processing combined data signals, and circuitry being adapted for transmitting inverse subband processed combined data signals with said spatial diversity means, is disclosed. Said apparatus can be exploited in the downlink transmission methods in the processing peer.

[0017] In an embodiment of the invention parallelism in the apparatus structure is introduced either in said circuitry being adapted for combining data or (and) in said circuitry being adapted for transmitting inverse subband processed combined data signals.

[0018] An apparatus having the functionality and architectural characteristics of both said apparatus in said third and fourth aspect of the invention can be defined.

Brief description of the drawings

[0019] Figure 1: (Up-link) communication setup with a composite peer (10), comprising of at least two transmitting terminals (20), each having at least one transmitting means (60) and a processing peer (30), comprising of at least one receiving terminal (40), having a spatial diversity receiving means (being represented here but not limited thereto by two spaced apart receiving means (80)). Said receiving terminal (40) at least performs subband processing (90). The top of figure 1 shows a concentrated scenario wherein said receiving terminal further performs inverse subband processing (110). The bottom of figure 1 shows a split scenario wherein the transmitting terminals further performs inverse subband processing (160).

[0020] Figure 2: (Down-link) communication setup with a composite peer (340), comprising of at least two receiving terminals (330), each having at least one receiving means (320) and a processing peer (230), having at least one transmitting terminal (240), having a spatial diversity transmitting means (being represented here but not limited hereto by two spaced apart transmitting means (220)). Said transmitting terminal (240) performs at least inverse subband processing (260). The top of figure 2 shows a concentrated scenario wherein said transmitting terminal (240) further performs subband processing (280). The bottom of figure 2 shows a split scenario wherein the receiving terminals performs subband processing (350).

[0021] Figure 3 shows the performance of an embodiment of the invention being uplink Least-Squares-OFDM/SDMA for one to four simultaenous users. The Bit Error Rate (BER) is shown as function of the Signal to Noise Ratio (SNR). Said embodiment can be exploited for an arbitrary number of users. In said embodiment OFDM is exploited as subband method, through exploiting of inverse fast Fourier transform and fast Fourier transform. Said determining of estimates of data signals from subband processed received data signals is in said embodiment based on Least-Squares methods.

[0022] Figure 4 shows the performance of pcSIC (per-carrier successive interference cancellation) for uplink OFDM/SDMA for one to four simultaneous users. The Bit Error Rate (BER) is shown as function of the Signal to Noise Ratio (SNR). Said embodiment can be exploited for an arbitrary number of users. Said determining of estimates of data signals from subband processed received data signals is in said embodiment based on determining an estimate of a selected data signal, in a Least-Squares sense, then modifying subband processed received data signals, and finally determining estimates of the remaining data signals, being all data signals except the selected data signal. For said remaining data signals said determining of estimates of a selected data signal can be exploited also. Said selection of data signals approach results in introducing an ordering in which estimates of data signals will be determined.

[0023] Figure 5: Error propagation with uplink SIC-OFDM/SDMA. The error count and the Signal Interference Ration (SIR) is shown as function of frequency. Said observation of the error propagation motivates the use of State Insertion methods.

[0024] Figure 6 shows the performance of pcSIC for uplink OFDM/SDMA with State Insertion for one to four simul-

taneous users. Said embodiment can be exploited for an arbitrary number of users. The Bit Error Rate (BER) is shown as function of the Signal to Noise Ration (SNR). Said determining of estimates of data signals from subband processed received data signals in said embodiment based on determining a plurality of estimates of a selected data signals, then accordingly modifying said subband processed received data signals, then determining estimates of at least one of the remaining data signals, being all the data signals except the selected data signal, and finally selecting one of said plurality of estimates of said selected data signal. Said selection of data signals approach results in introducing an ordering in which estimates of data signals will be determined. The amount of estimates to be determined per data signal can differ and can possibly be one for some data signals.

[0025] Figure 7 shows the performance of downlink OFDM/SDMA for one to four simultaneous users. Said embodiment can be exploited for an arbitrary number of users. The Bit Error Rate (BER) is shown as function of the Signal toNoise Ratio (SNR).

[0026] Figure 8 shows the performance of uplink OFDM/SDMA for different coherence grouping factors. The Bit Error Rate (BER) is shown as function of the Signal to Noise Ration (SNR). In said embodiment the initialization, being the determining of relations between data signals and subband processed received data signals, is performed on a set-by-set basis. Said coherence grouping factor denote the amount of subbands in a set.

[0027] Figure 9 presents the operation count (initialization and processing) of Least-Squares-OFDM/SDMA.

[0028] Figure 10 present the operation count (initialization and processing) of pcSIC (per-carrier successive interference cancellation)-OFDM/SDMA.

[0029] Figure 11 presents the additional operation count for State Insertion (initialization and processing).

Detailed description of the invention

[0030] The invention is not limited to the detailed description of the invention found below.

[0031] The invention concerns (wireless) communication between terminals (20) (40) (330) (240). We can logically group the terminals on each side of the communication and call them peers (340) (230) (10) (30). Each peer can embody one or more terminals, whereby at least one peer embodies more than one terminal. The focus of the invention is thus multi-user. The peers can be transmitting and/or receiving information. For example, the peers can communicate in a half-duplex, being either transmitting or receiving at one time instance or a full duplex fashion, being substantially simultaneously transmitting and receiving.

[0032] The invention introduces Space Division Multiple Access (SDMA) techniques for systems making use of subband processing, and thus fits in a multi-carrier approach. In the invention, at least one of the communication peers (30) (230) consists of terminal(s) (40) (240) disposing of transmitting and/or receiving means (80) (220) that are able to provide different spatial samples of the transmitted and/or received signals. We will call these transmission and/or receiving means spatial diversity means. We will call the peer(s) disposing of said spatial diversity means the processing peer(s) (30)(230). Said processing peer communicates with at least two terminals at the opposite peer (10) (340), which can operate at least partially simultaneously and the communicated signals' spectra can at least partially overlap. Note that Frequency Division Multiple Access techniques rely on signals spectra being non-overlapping while Time Division Multiple Access techniques rely on communicating signals in different time slots thus not simultaneously. We will call this opposite peer(s) consisting of at least two terminals (20) (330) using the same frequencies at the same time, the composite peer(s) (10)(340). The invention concerns (wireless) communication between terminals whereby at least the processing peer(s) (30) (230) disposes of subband processing means.

[0033] The communication between the composite peer and the processing peer can consist of uplink (Figure 1) and downlink (Figure 2) transmissions. With uplink transmission is meant a transmission whereby the composite peer transmits data signals and the processing peer receives data signals. With downlink transmission is meant a transmission whereby the processing peer transmits data signals and the composite peer receives data signals. The uplink and downlink transmissions can either be simultaneous (full duplex) (for example using different frequency bands), or they can operate in a time-duplex fashion (half duplex)(for example using the same frequency band), or any other configuration.

[0034] (Wireless) transmission of data or a digital signal from a transmitting to a receiving circuit requires digital to analog conversion in the transmission circuit and analog to digital conversion in the receiving circuit. In the further description it is assumed that the apparatus in the communication set-up have transmission and receiving means, also denoted front-end, incorporating these analog to digital and digital to analog conversion means including amplification or signal level gain control and realizing the conversion of the RF signal to the required baseband signal and vice versa. A front-end can comprise of amplifiers, filters and mixers (down converters). As such in the text all signals are represented as a sequence of samples (digital representation), thereby assuming that the above mentioned conversion also takes place. Said assumption does not limit the scope of the invention though. Communication of a data or a digital signal is thus symbolized as transmitting and receiving of a sequence of (discrete) samples. Prior to transmission, the information contained in the data signals can be fed to one or more carriers or pulse-trains by mapping said data signals

to symbols which consequently modulate the phase and/or amplitude of the carrier(s)or pulse-trains (e.g. using QAM or QPSK modulation). Said symbols belong to a finite set, which is called the transmitting alphabet. The signals resulting after performing modulation and/or front-end operations on said data signals, are called transformed data signals, to be transmitted further.

[0035] After reception by the receiving means, the information contained in the received signals is retrieved by transformation and estimation processes. These transformation and estimation processes can include but should not include demodulation, subband processing, decoding, equalization. After said estimation and transformation processes, received data signals are obtained consisting of symbols belonging to a finite set, which is called the receiving alphabet. The receiving alphabet is preferably equal to the transmitting alphabet.

[0036] The invention can further exploit methods and means for measuring the channel impulse responses between the transmission and/or reception means of the individual terminals at the composite peer on the one hand, and the spatial diversity means of the processing peer on the other hand. The channel impulse responses measurement can be either obtained on basis of an uplink transmission and/or on basis of a downlink transmission. The thus measured channel impulse responses can be used by the processing peer and/or composite peer in uplink transmissions and/or in downlink transmissions. The invention can further exploits methods for determining the signal power of received data signals and methods for determining the interference ratio of data signals.

[0037] The spatial diversity means ensures the reception or transmission of distinct spatial samples of the same signal. This set of distinct spatial samples of the same signal is called a spatial diversity sample. In an embodiment, spatial diversity means embody separate antennas. In this embodiment, the multiple antennas belonging to one terminal can be placed spatially apart (as in Figure 1 and 2), or they can use a different polarization. The multiple antennas belonging to one terminal are sometimes collectively called an antenna array. The invention is maximally efficient if the distinct samples of the spatial diversity sample are sufficiently uncorrelated. In an embodiment, the sufficiently uncorrelated samples are achieved by placing different antennas apart over a sufficiently large distance. For example, the distance between different antennas can be chosen to be half a wavelength of the carrier frequency at which the communication takes place. Spatial diversity samples are thus different from each other due to the different spatial trajectory from the transmitting means to their respective receiving means or vice versa. Alternatively said spatial diversity samples are different from each other due to the different polarization of their respective receiving or transmitting means.

[0038] The methods described rely on the fact that at least the processing peer performs subband processing, called SP in the sequel, in the uplink mode (Figure 1), and inverse subband processing, called ISP in the sequel, in the downlink mode (Figure 2). Furthermore, in the uplink mode ISP takes place either in the composite peer prior to transmission (see Figure 1 bottom) or in the processing peer after SP (see Figure 1 top). In the downlink mode, SP takes place either in the composite peer after reception (see Figure 2 bottom) or in the processing peer before ISP (see Figure 2 top). The scenarios where both ISP and SP are in either transmission direction carried out in the processing peer, are called concentrated scenarios. The remaining scenarios, i.e. where ISP and SP are carried out in different peers in either transmission direction, are called split scenarios.

[0039] The uplink transmission methods can be formalized as a first aspect of the invention being methods of data signals (50) from at least two transmitting terminals (20) with each at least one transmitting means (60) to at least one receiving terminal (40) with a spatial diversity receiving means (80) comprising the following steps: (first step) transmitting from said transmitting terminals (20) transformed data signals (70), being transformed versions of said data signals; (second step) receiving on said spatial diversity means (80) received data signals being at least function of at least two of said (transmitted) transformed data signals (70); (third step) subband processing (90) of at least two of said received data signals in said receiving terminal (40); and (fourth step) determining estimates of said data signals (120) from (the obtained) subband processed received data signals (140) in said receiving terminal.

[0040] Said transmitting of said transformed data signals can be substantially simultaneously. The spectra of said transformed data signals can be at least partly overlapping.

[0041] In the uplink split scenario said transformation of said data signals (50) to transformed data signals (70) comprises inverse subband processing (160). In the uplink concentrated scenario said determining (150) of estimates of said data signals from (the obtained) subband processed received data signals in said receiving terminal comprises the following steps: (first step) determining (100) intermediate estimates of said data signals (130) from said subband processed received data signals in said receiving terminal; (second step) obtaining said estimates of said data signals (120) by inverse subband processing (110) said intermediate estimates.

[0042] The downlink transmission method can be formalized as a second aspect of the invention being methods of transmitting data signals (200) from at least one transmitting terminal (240) with a spatial diversity transmitting means (220) to at least two receiving terminals (330)with at least one receiving means (320) comprising the following steps: (first step) determining (250) combined data signals (300) in said transmitting terminal, said combined data signals being transformed versions of said data signals; (second step) inverse subband processing (260) said combined data signals; (third step) transmitting with said spatial diversity means (220) (the obtained) inverse subband processed

combined data signals; (fourth step) receiving on at least one of said receiving means (320) of at least one receiving terminal (330) inverse subband processed received data signals; (fifth step) determining estimates of said data signals from said inverse subband processed received data signals.

[0043] Said transmitting can be substantially simultaneously. The spectra of said (transmitted) inverse subband processed combined data signals can be at least partly overlapping;

[0044] In the downlink split scenario said determining of said estimates of said data signals in said receiving terminals comprises subband processing (350). In the downlink concentrated scenario determining combined data signals in said transmitting terminal comprises the following steps: (first step) determining intermediate combined data signals (290) by subband processing (280) said data signals; (second step) determining (270) said combined data signals from said intermediate combined data signals.

[0045] It must be mentioned that said transmission methods intend to transmit data signals from one peer to another peer but that due to transmission conditions in fact only estimates of said data signals can be obtained in the receiving peer. Said transmission methods naturally intend to be such that said estimates of said data signals are approximating said data signals as close as technically possible.

[0046] It is a characteristic of the invention that said transmission methods are not a straightforward concatenation of a Space Division Multiple Access technique and a multi-carrier modulation method. Indeed a straightforward concatenation would be to modulate in the transmitting terminal with such a multi-carrier method, then transmit in a Space Division Multiple Access setting and at the receiving terminal combine the different signals received on the spatial diversity means into a combined signal and then demodulate the combined signal. In the split scenario of the invention in the transmitting terminal the (to be transmitted) transformed signal is also modulated with a multi-carrier method and there is a transmission in a Space Division Multiple Access setting but in the receiving terminal first the different received signals on the spatial diversity means are demodulated and only then the demodulated received signals are combined. In the concentrated scenario both modulation and demodulating is concentrated in the receiving peer in the uplink and in the transmitting peer in the downlink. Note that here with modulation and demodulation subband and inverse subband processing is meant and not standard modulation.

[0047] In an embodiment of both the first and second aspect of the invention implements a multicarrier modulation technique. An example of such a multicarrier modulation technique uses IFFT as ISP and FFT as SP, and the modulation technique is called Orthogonal Frequency Multiplexing (OFDM) modulation. It can be stated that in said uplink transmission method that said subband processing is orthogonal frequency division demultiplexing. It can also be stated that in said uplink transmission method that said inverse subband processing is an orthogonal frequency division multiplexing. It can also be stated that in said downlink transmission method that said subband processing is orthogonal frequency division demultiplexing. It can also be stated that in said downlink transmission method that said inverse subband processing is orthogonal frequency division multiplexing.

[0048] In concentrated scenarios, the processing that is carried out in the processing peer on samples between SP (90) (280) and ISP (110) (260) is called subband domain processing (270)(100). In split scenarios, the processing that is carried out prior to ISP (160) (260) in the transmitting terminal(s) and after SP (90) (350) in the receiving terminal (s), is called subband domain processing (eg. (250)). With before is meant coming earlier in time during the transmission or the reception, and with after is meant coming later in time during the transmission or the reception. In concentrated scenarios, the signals (130) (140) (290) (300) between the SP and the ISP are called signals in subband domain representation. In split scenarios, the signals (50) (300) (200) before the ISP in the transmitting terminal(s) and the signals (360)(140)(120) after the SP in the receiving terminal(s) are called signals in a subband domain representation.

[0049] In an embodiment, the subband processing consists of Fast Fourier Transform (FFT) processing and the inverse subband processing consists of Inverse Fast Fourier Transform Processing. By FFT processing is meant taking the Fast Fourier Transform of a signal. By Inverse FFT processing is meant taking the Inverse Fast Fourier Transform of a signal.

[0050] In the invention the transmitted sequence is divided in data subsequences prior to transmission. Said data subsequences correspond to subsequences that will be processed as one block by the subband processing means. In case of multipath conditions, a guard interval containing a cyclic prefix or postfix is inserted between each pair of data subsequences in the transmitting terminal(s). If multipath propagation conditions are experienced in the wireless communication resulting in the reception of non-negligible echoes of the transmitted signal and the subband processing means consist of (an) FFT and/or IFFT operation(s), this guard introduction results in the equivalence between convolution of the time-domain data signals and the time-domain channel response with multiplication of the frequency-domain data-signals and the frequency-domain channel response. The insertion of said guard intervals can occur in both concentrated and split scenarios. One can thus state that in an embodiment of the invention in a split scenario, the transmitting terminal(s) insert guard intervals containing a cyclic prefix or postfix between each pair of data subsequences after performing ISP on the data subsequences and before transmitting the data subsequences. In another embodiment of the invention in a concentrated scenario, said guard intervals are inserted in the transmitted sequence between each pair of data subsequences without performing ISP on said data subblocks in the transmitting terminal

(s). This can be formalized as follows by stating that in said uplink transmission methods said transformation of said data signals to transmitted data signals further comprises of guard interval introduction. Said guard interval introduction can equally well be applied in the downlink transmission methods. Alternatively overlap and save techniques can be exploited also.

[0051] The terminal(s) disposing of the spatial diversity means (thus in the processing peer) dispose(s) of SP and/or ISP means that enable subband processing of the distinct samples of the spatial diversity sample. Also, it disposes of means for combinatory processing. By combinatory processing means is meant means that process data coming from subbands of the distinct samples in the spatial diversity sample. In said combinatory processing means, different techniques can be applied to retrieve or estimate the data coming from the different distinct terminals or to combine the data to be transmitted to distinct terminals. The invention discloses methods for performing said combinatory processing, both for uplink transmission and for downlink transmission.

[0052] Combinatory processing in the uplink relates to a communication situation whereby the peer disposing of spatial diversity means, which is called the processing peer, is receiving signals from the composite peer, which embodies different terminals transmitting (at least partially simultaneous) transformed data signals (having at least partially overlapping spectra). With said determining of estimates of said data signals (120) from said subband processed received data signals (140) in said receiving terminal in said uplink transmission method is meant said combinatory processing.

[0053] Combinatory processing in the downlink relates to a communication situation whereby the peer disposing of spatial diversity means, which is called the processing peer, is transmitting signals to the composite peer, which embodies different terminals transmitting (at least partially simultaneous) so-called inverse subband processed combined data signals (having at least partially overlapping spectra). With determining (250) combined data signals (300) in said transmitting terminal in said downlink transmission method is meant said combinatory processing.

[0054] In the invention further the following notations are used: x is used for a transmitted data sample. The notation y is used for a received data sample. The notation n is used for a noise sample. The notation X is used for a transmitted data sample matrix. The notation Y is used for a received data sample matrix. The notation N is used for a noise sample matrix. The notation greek symbol sigma is used for the variance of the noise. The notation h(t) is used for the channel impulse response represented in the time domain. The notation h[s] is used for the channel impulse response represented in the frequency domain. The array index s (going from 1 to S) refers to the specific subband to which a sample or a channel impulse response corresponds. The notation S is used for the total number of subbands that are processed by the subband processing means. The superscript index u (going from 1 to U) refers to the individual terminal of the composite peer by which the data signal was sent in the uplink mode, or for which the data signal is intended in the downlink mode. The notation U is used for the number of simultaneous terminals of the same subbands in the composite peer. The subscript index a (going from 1 to A) refers to one specific spatial sample of the spatial diversity sample in the processing peer. The notation A is used for the number of distinct samples in the spatial diversity sample in the processing peer The notation e is used for an equalizer coefficient. The notation E is used for an equalizer coefficient matrix. The notation greek epsilon is used for the stochastical expectation operator. A ~ on top of a symbol indicates a soft estimation of the symbol. With a soft estimation is meant an estimation that is not necessarily contained in the receiving alphabet. A bar on top of a symbol indicates a hard estimation of the symbol. With a hard estimation is meant an estimation that is equal to a symbol contained in the receiving alphabet.

[0055] In the invention decision methods are exploited. Said decision methods obtain one or more intermediate hard estimates, on basis of a single soft estimate. In an embodiment of the invention, said decision methods obtain one intermediate hard estimate by determining that signal from the receiving alphabet that has the smallest distance to the soft estimate. In another embodiment of the invention, said decision methods obtain multiple intermediate hard estimates by determining those signals from the receiving alphabet that have the smallest distance to the soft estimate.

[0056] In the invention selection methods are exploited. Said selection methods obtain hard estimates for a specific data symbol on basis of intermediate hard estimates of said specific data symbol. In an embodiment of the invention, there is only one intermediate hard estimate for a specific data signal/symbol and the hard estimate is equal to the intermediate hard estimate. In another embodiment, there are several intermediate hard estimates for at least one data signal. From multiple intermediate hard estimates, one intermediate hard estimate for one specific data signal is selected as the hard estimate on basis of a probability criterion.

[0057] In the invention recombiner methods are exploited. Said recombiner methods obtain recombined spatial diversity samples, on basis of a hard or intermediate hard estimate, by calculating the spatial diversity samples that would have been received if the data symbol corresponding to said hard or intermediate hard estimate would have been transmitted.

[0058] In the invention modifier methods are exploited. Said modifier methods obtain modified spatial diversity samples, by applying the following steps. First, they obtain recombined spatial diversity samples by applying recombiner methods based on previously obtained hard or intermediate hard estimates. Secondly, they obtain modified spatial diversity samples by exploiting the recombined spatial diversity samples and the original spatial diversity samples.

[0059] Said modifier methods are exploited in the uplink transmission methods, denoted successive interference cancellation methods, wherein said determining of said estimates of said data signals from said subband processed received data signals in said receiving terminal further comprising for at least one data signal the following steps: (first step) selecting from said data signals a selected data signal; (second step) determining an estimate of said selected data signal from said subband processed received data signals; (third step) modifying said subband processed received data signals based on said estimate of said selected data signal via a modifier method; and (fourth step) determining estimates of said remaining data signals from said modified subband processed received data signals. Note that said selection of a selected data signals is just determining for which signal said method will be applied. Said selection should not be confused with the selection methods described above.

[0060] Said modifier methods are also exploited in the uplink transmission methods, denoted state insertion methods, wherein said determining of said estimates of said data signals from said subband processed received data signals in said receiving terminal further comprising for at least one data signal the following steps: (first step)selecting from said data signals a selected data signal; (second step) determining a plurality of estimates of said selected data signal from said subband processed received data signals; (third step) determining a plurality of modified subband processed received data signals, each of said modified subband processed received data signals being based on one of said estimates of said selected data signal, via a modifier method; (fourth step) determining a plurality of estimates of at least one of said remaining data signals from said plurality of modified subband processed received data signals; and (fifth step) thereafter selecting one of said estimates of said selected data signal (by a selection method). Note that said selection of a selected data signals is just determining for which signal said method will be applied. Said selection should not be confused with the selection methods described above.

[0061] It can be mentioned that said estimate of said selected data signal in said successive interference cancellation methods can be considered to be a hard estimate. Said plurality of said estimates of said selected data signal in said state insertion methods can be considered as intermediate hard estimates. Said plurality of estimates of said remaining data signals in both said methods can be either (intermediate) hard or soft estimates.

[0062] In the invention, methods for coherence grouping of subbands is presented. Said coherence grouping methods reduce the initialization effort, and can be used both in uplink and downlink transmissions. Said coherence grouping methods partition the S subbands in groups of adjacent subbands, each group i consisting of G_l subbands. The initialization computations are in said coherence grouping methods performed only once for each subgroup, instead of for each subband separately. Said coherence grouping methods do not affect the performances of the combinatory processing methods, provided all subbands in a subgroup experience sufficiently correlated channel impulse responses. Therefore, the numbers G_l are limited by the communication situation in order for the invention to be maximally efficient. For the initialization effort, said coherence grouping method results in a reduction of the computation complexity with a factor G, whereby G is the average of the numbers G_l of the subbands in the subgroups. In an embodiment, the communication is based on OFDM transmission and the numbers G_l are all chosen equal to a fixed part of the coherence bandwidth divided by the spacing between the carriers. The coherence bandwidth of the channel is the bandwidth over which the channel response is correlated. On a multipath propagation channel, said coherence bandwidth is inversely proportional to the relative delay of the echoes on the channel. In another embodiment, the numbers G_l are calculated on basis of a channel impulse response measurement, more specifically from the gradient of this channel impulse response. More formalized one can state that said determining of said estimates of said data signals in said receiving terminal comprises two steps. The first step is the initialization step wherein relations between said data signals and subband processed received data signals are determined. In the second step being the actual combinatory processing said relations between said data signals and said subband processed received data signals are exploited for determining said data signals. The coherence grouping method is then characterized by stating said subbands are grouped into sets, at least one set comprising of at least two subbands and that said initialization step is performed on a set-by-set basis.

[0063] In the invention, methods for combinatory processing for uplink communication are presented. Said combinatory processing for uplink communication methods obtain in the receiving terminal estimates for datasignals transmitted from one or more terminals in the composite peer, on basis of the subband processed spatial diversity samples, which can also be denoted subband processed received data signals.

[0064] In the invention, methods for per-subband combinatory processing for uplink communication are presented. Said per-subband combinatory processing for uplink communication methods obtain in the receiving terminal estimates for data signal(s) transmitted from one or more terminals in the composite peer and in one specific subband, on basis of the subband processed spatial diversity samples in that one specific subband, which can also be denoted subband processed received data signals in that one specific subband. Said per-subband combinatory processing methods can be formalized as methods in which said determining of estimates of said data signals in said receiving terminal is performed on a subband by subband basis.

[0065] In an embodiment of the invention, said methods for per-subband combinatory processing for uplink communication obtain estimates for data signal(s) from at least one terminal in the composite peer and in one specific subband,

by applying the following steps. First, they obtain soft estimates for the data signal(s) from each of these terminals and in that specific subband, by applying per-subband scalar combinatory processing for uplink communication methods. Secondly, they obtain estimates by applying decision methods on each of said soft estimates.

[0066] In the invention, methods for per-subband scalar combinatory processing for uplink communication are presented. Said per-subband scalar combinatory processing for uplink communication methods obtain soft estimates for data signal(s) in one subband and from one terminal in the composite peer, on basis of spatial diversity samples or modified spatial diversity samples in that subband. In an embodiment of the invention, said methods for per subband scalar combinatory processing for uplink communication obtain soft estimates for data signal(s) in one subband and from one terminal in the composite peer with linear methods. In this embodiment, the estimates of the data signal(s) transmitted by the terminal(s) of the composite peer $\tilde{x}^{U}$ [s] are calculated by linearly combining the single corresponding carrier signals or subbands received on the different antennas with the equalizer coefficients E[s], following Formula 1 below.

$$\underbrace{\begin{bmatrix} \tilde{x}^{1}[s] \\ \vdots \\ \tilde{x}^{U}[s] \end{bmatrix}}_{\tilde{X}[s]} = \underbrace{\begin{bmatrix} e_1^1[s] & \dots & e_A^1[s] \\ \vdots & & \vdots \\ e_1^U[s] & \dots & e_A^U[s] \end{bmatrix}}_{E[s]} \cdot \begin{bmatrix} y_1[s] \\ \vdots \\ y_A[s] \end{bmatrix} \qquad \text{Formula 1}$$

[0067] In an embodiment, the linear estimation is performed on basis of least-squares (LS) methods. In this embodiment of the invention, said E[s] is calculated to minimize the expectations given in Formula 2. For a given noise energy (sigma squared) and conditions on x given by Formula 3, E[s] obeys the U sets of linear equations of Formula 4, wherein the superscript H denotes the Hermitian transpose.

$$\mathcal{E}\left\{\left(x^{u}[s]-\tilde{x}^{u}[s]\right)^{*}\left(x^{u}[s]-\tilde{x}^{u}[s]\right)\right\} \qquad \text{Formula 2}$$

$$\mathcal{E}\left\{x^{u}[s]^{*}\tilde{x}^{u}[s]\right\}=1 \qquad \text{Formula 3}$$

$$\left[H[s]H[s]^{H}+\sigma^{2}\, I_{A\times A}\right]E^{H}-H[s]=0 \qquad \text{Formula 4}$$

[0068] In an embodiment of the invention, said methods for per subband scalar combinatory processing for uplink communication obtain soft estimates for data signal(s) in one subband and from one terminal in the composite peer by linear zero forcing (ZF) methods. In this embodiment of the invention, said E[s] is calculated to maximally annihilate the channel distortion without taking the noise energy into account, in a so-called zero-forcing way. In this embodiment E[s] obeys the U sets of linear equations of Formula 5.

$$\left[H[s]^{H}\right]E^{H}-I_{U\times U}=0 \qquad \text{Formula 5}$$

[0069] In an embodiment of the invention, said methods for per subband combinatory processing for uplink commu-

nication obtain soft estimates for data signal(s) in one subband and from one terminal in the composite peer with non-linear methods such as for example maximum likelihood symbol estimation (MLSE).

[0070] In the invention, methods for combinatory processing for downlink communication are presented, called downlink combinatory processing methods. Said downlink combinatory processing methods are carried out in the processing peer to facilitate the estimation in the composite peer of the transmitted data signals from the processing peer. Said downlink combinatory processing methods produce a spatial diversity sample, on basis of at least two data signals, resulting in combined data signals. Said combined data signals then undergo ISP, and are afterwards transmitted by the spatial diversity means, resulting in transmitted data signals, the spectra of said transmitted data signals being at least partly overlapping. The transmitted data signals are then transmitted on the channel. Said combinatory processing can be described as a step for determining combined data signals in said transmitting terminal, said combined data signals being transformed versions of said data signals. Thereafter inverse subband processing of said combined data signals is performed, followed by transmitting with said spatial diversity means said inverse subband processed combined data signals, the spectra of said transmitted inverse subband processed combined data signals being at least partly overlapping.

[0071] In an embodiment of the invention, said methods for combinatory processing for downlink communication apply per-subband combinatory processing for downlink communication in at least one subband. In this embodiment, said downlink combinatory processing methods produce a combined data signal in one subband, on basis of the data signals in that specific subband. This can be described as determining of combined data signals in said transmitting terminal on a subband by subband basis.

[0072] In an embodiment of the invention, said downlink combinatory processing methods calculate said combined data symbol, called Y[s] by linearly combining the data signals, called X[s], with a precompensation matrix $E_{PRE}[s]$. In this embodiment, the Formula 6 holds:

$$X[s] = H^T[s] \, . \underbrace{E_{PRE}[s] \, . \, Y[s]}_{Y_{TR}[s]} + N''[s] \qquad \text{Formula 6}$$

[0073] In an embodiment, U = A and $E_{PRE}[s]$ is the inverse of the channel matrix $H^{T(-1)}[s]$.

[0074] In another embodiment, U < A and $E_{PRE}[s]$ is the pseudo-inverse of the channel matrix such that Formula 7 holds:

$$H^T[s]. E_{PRE}[s] = I_{U \times U} \qquad \text{Formula 7}$$

[0075] In an embodiment of the invention, said methods for combinatory processing for downlink communication take into account non-idealities in the analogue front-end(s) in the transmitting- and or receiving terminal(s).

[0076] In an embodiment of the invention, said methods for combinatory processing for downlink communication avoid distortion in the analogue front-end(s).

[0077] In an embodiment of the invention, said precompensation matrix comprises nulls for one or more data signals on one or more specific subbands. In this embodiment, efficient and effective power usage can be obtained.

[0078] In an embodiment of the invention, as well the terminals in the processing peer as the terminals in the composite peer dispose of spatial diversity means and of subband and combinatory processing means, and both the processing and the composite peers embody at least two terminals transmitting data signals at least partially simultaneously whereby these transmitted data signals have at least partially overlapping spectra. In this embodiment, both the uplink and the downlink combinatory processing methods can be used in either transmission direction between two peers.

[0079] In a preferred embodiment, the processing peer is the basestation of a star-configuration network, which can be connected to the backbone of a wired network. Different terminals communicate simultaneously in the same frequency band with the basestation. The terminals only need a single front-end, and a plain OFDM-modem (i.e. without SDMA processing capabilities). This embodiment can work in a multipath fading environment, for applications such as for example Wireless Local Area Networks. This embodiment can work in the 5-6 GHz band, and can achieve a network capacity of 155 Mbps and above. In this embodiment, the spatial diversity means consisting of multiple transmitting and/or receiving means are concentrated in the basestation, thus reducing the overall hardware complexity and cost of the configuration. In this embodiment, the spatial diversity means comprises of multiple antennas spaced half a wavelength apart. Also, the peers communicate on basis of OFDM as a modulation technique. The ISP and SP are

performed on basis of IDFT (Inverse Discrete Fourier Transform) and DFT (Discrete Fourier Transform) processing respectively, and the subbands can be referred to as carriers. A guard interval containing a cyclic prefix is inserted between OFDM symbols at the transmitter side. SDMA is then applied in the frequency-domain. In addition to the aggregate of the advantages of both OFDM and SDMA, this approach results in simpler exploitation of spatial diversity compared to time-domain methods. For the estimation algorithms of the OFDM/SDMA systems, known least-squares (LS) or maximum likelihood symbol estimation (MLSE) algorithms can be applied. Also a novel class of uplink OFDM/SDMA methods with improved performance and still low complexity can be applied. These methods implement successive interference cancellation per carrier, state insertion and coherence grouping.

[0080] To illustrate the performance, the algorithms are applied to a 100 Mbps OFDM/SDMA WLAN. It has a 4-antenna basestation that separates up to 4 simultaneous users by SDMA. Each of these transmit 256-OFDM symbols at a datarate of 25 Mbps with QPSK. The guard interval is designed to comprise all the echoes received on the multipath propagation channel, so that the channel convolution becomes cyclic after removal of the guard interval. Thus, in the frequency-domain it becomes equivalent to multiplication with the Fourier transform of the channel, $h^U{}_a[s]$. Measures are taken to synchronize the different users communicating simultaneously in the same band. Under these conditions, the received data Y[s] can be written as in Formula 8. It is possible under the given circumstances to calculate an estimation $\hat{x}^u$ [s] of the data sequences X^U[s] on a carrier per carrier basis. This per carrier estimation greatly simplifies the SDMA processing, and it allows a profound parallellization of this processing. Indeed an apparatus for performing SDMA/OFDM can comprise of a plurality of circuits being adapted for determining estimates of data signals based on part of the subbands of the subband processed received data signals, preferably one subband per circuit. Subsequently to the per carrier estimation, a decision on which element of the transmitting alphabet is nearest to the signal(s) resulting from the per carrier estimation. Said decision results in hard estimates $\tilde{x}^U$.

$$\underbrace{\begin{bmatrix} y_1[s] \\ \vdots \\ y_A[s] \end{bmatrix}}_{Y[s]} = \underbrace{\begin{bmatrix} h_1^1[s] \ldots h_1^U[s] \\ \vdots \quad\quad \vdots \\ h_A^1[s] \ldots h_A^U[s] \end{bmatrix}}_{H[s]} \cdot \underbrace{\begin{bmatrix} x^1[s] \\ \vdots \\ x^U[s] \end{bmatrix}}_{X[s]} + \underbrace{\begin{bmatrix} n_1[s] \\ \vdots \\ n_A[s] \end{bmatrix}}_{N[s]} \qquad \text{Formula 8}$$

[0081] The per carrier SDMA processing in the uplink can be performed in a linear fashion. In this case, the estimates of the data signals transmitted by the terminals $\tilde{x}^U$ [s] are calculated by linearly combining the single corresponding carrier signals received on the different antennas with the equalizer coefficients E[s], following Formula 1.

[0082] Said E[s] is calculated to minimize the expectations given in Formula 2. For a given noise energy (sigma squared) and conditions on x given by Formula 3, E[s] obeys the U sets of linear equations of Formula 4, wherein the superscript H denotes the Hermitian transpose.

[0083] Subsequently to equalization, the soft estimates are fed into a slicer, which determine the nearest constellation points with a decision method. This results in the hard estimates $\tilde{x}^U$. The equalizer coefficients E[s] could alternatively be calculated to maximally annihilate the channel distortion without taking the noise energy into account, in a so-called zero-forcing way. In this method, E[s] obeys the U sets of linear equations, given by Formula 5.

[0084] The performance of the system is evaluated by simulation with realistic channel data obtained from ray-tracing. The resulting curves show the bit error rate (BER) as a function of the received signal-to-noise ratio per bit (SNR).

[0085] Figure 3 shows the performance of LS-OFDM/SDMA for one to four simultaneous users. As a reference, the dashed curve gives the performance of a single-user single-antenna 100 Mbps plain OFDM link. An important observation is that the four-antenna four-user LS-OFDM/SDMA system outperforms plain OFDM. This demonstrates that a bandwidth re-use factor four is achievable without any performance penalty.

[0086] To evaluate the implementation complexity, we determine the total number of operations needed to execute the LS-OFDM/SDMA algorithm. The overall functionality can be separated into initialization and processing. During initialization, the equalizers E_{LS}[s] are calculated from Formula 4 by Gaussian elimination with multiple right-hand sides. During processing, equalization and slicing have to be performed, which respectively correspond to a matrix multiplication and a set of comparators. Note that processing is done continuously, at the symbolrate, opposed to the initialization which is only calculated once. Figure 9 summarizes the approximative number of multiplications, additions and data transfers needed for the execution of both phases, per sub-carrier. The number of data transfers is an indicative number for the amount of memory/register transfers. Since the architecture has not been determined yet, the allocation of the data transfers to memory or register banks is an open issue. However, it is important to stress that data transfers often are the implementation bottleneck. As an example, the four-user four-antenna system from would require 72

kflops and 200kdata transfers during initialization and 400 Mflops/sec and a data transfer bandwidth of 1.2 Gwords/sec during processing.

[0087] In an OFDM/SDMA system, for certain subcarriers one or more users may be completely buried in multi-user interference (MUI) or have highly correlated channel vectors. Obviously, these users will suffer from residual MUI and noise after linear equalization. To mitigate this effect, in the invention Successive Interference Cancellation (SIC) is used, preferably Per Carrier (pcSIC). Opposed to LS-OFDM/SDMA, this technique does not estimate all users simultaneously, but does this successively, preferably on a certain subcarrier. As such the MUI originating from users that have been detected already, can be removed (thereby modifying the original signal carrier). This technique relies on feedback of intermediate hard estimates and is thus non-linear. Note also that pcSIC-OFDM/SDMA -since it works preferably on a per-carrier basis- elegantly exploits frequency diversity, which is another example of synergy from the OFDM/SDMA combination. Further per-carrier successive interference cancellation (pcSIC) is presented, although the invention is not limited hereto. On each subcarrier n, the detection order is first determined, preferably according to the received signal power. We assume that user 1 to user U are properly ordered. Successively, each user's soft estimate is calculated by least-squares linear combining according to Formula 9.

$$\tilde{x}^u[s] = \underbrace{\left[e_1^u[s] \cdots e_A^u[s]\right]}_{E_{LS}^u[s]} \cdot \begin{bmatrix} y_1^u[s] \\ \vdots \\ y_A^u[s] \end{bmatrix} \qquad \text{Formula 9}$$

$$H^u[s] = \begin{bmatrix} h_1^u[s] & \ldots & h_1^U[s] \\ \vdots & & \vdots \\ h_A^u[s] & \ldots & h_A^U[s] \end{bmatrix} \qquad \text{Formula 10}$$

[0088] In this equation $E_{LS}[s]$ is found as in Formula 4 except that for each iteration u, H[s] is replaced by Formula 10 with the $y^1_a[s]$ equal to the initial $y_a[s]$ from Formula 8. Afterwards, the MUI originating from user u is reconstructed (in general called recombining) and subtracted (thereby modifying) from the residual MUI as in Formula 11 with $\bar{x}^U[s]$ the hard intermediate estimate of user u after slicing (being a decision method). In the plain pcSIC algorithm, there is only one intermediate hard estimate for each data signal on each carrier, which will consequently be selected as the hard estimate (being a selection method).

$$\begin{bmatrix} y_1^{u+1}[s] \\ \vdots \\ y_A^{u+1}[s] \end{bmatrix} = \begin{bmatrix} y_1^u[s] \\ \vdots \\ y_A^u[s] \end{bmatrix} - \begin{bmatrix} h_1^u[s] \\ \vdots \\ h_A^u[s] \end{bmatrix} \bar{x}^u[s] \qquad \text{Formula 11}$$

[0089] Successive Interference Cancellation (SIC) can be described as an uplink transmission method wherein said determining of said estimates of said data signals from said subband processed received data signals in said receiving terminal for at least one data signal (but not limited to one) the following steps: (first step) selecting from said data signals a selected data signal. This selecting step can for instance be but is not limited to selecting the first data signal from the data signals being ordered according to received signal power. (second step) determining an estimate of said selected data signal from said subband processed received data signals, for instance but not limited to linear combining of the subband processed received data signals including slicing. (third step) modifying said subband processed received data signals based on said estimate of said selected data signal, for instance but not limited to recombining

and subtraction according to Formula 11, such that modified subband processed received data signals are obtained. (fourth step) finally estimates of said remaining data signals from said modified subband processed received data signals are determined, possibly by applying the same steps successively. Note that subband can also be denoted carriers.

[0090] In an embodiment of the successive interference cancellation uplink transmission method said determining of estimates (all steps) is performed on a subband by subband basis.

[0091] Figure 4 shows the performance of pcSIC-OFDM/SDMA. Again, the BER over SNR curves are given for one to four users. As a reference, the dashed curve gives the performance of single-user single-antenna 100 Mbps plain OFDM and the thin curves give the performance of LS-OFDM/SDMA. It shows that pcSIC-OFDM/SDMA yields a performance improvement compared to LS-OFDM/SDMA that increases with an increasing number of users. To be specific, for a BER of 0.001 and for four simultaneous users, we observe a 5 dB gain. For a complexity estimation of the pcSIC algorithm for OFDM/SDMA, we also discern between initialization and processing. During initialization, the cancellation order is determined and the equalizers are calculated. Since the channel matrix is different for each user u the latter would require U distinct Gaussian eliminations. However, by exploiting structural relationships between the $H^u[s]$'s, we managed to reduce the required number of operations with $O(A^2\ U^2)$. During processing, reconstruction (recombination) and subtraction (modification) have to be performed in addition to equalization and slicing. The total approximative number of operations for both phases of the pcSIC-OFDM/SDMA algorithm are given in Figure 10 . More specific, the four-user four-antenna system would require respectively 170kflops and 250kdata-transfers in the initialization phase and 700Mflops/sec and a data-transfer bandwidth of 1.8Gwords/sec.

[0092] A potential weakness of (pc)SIC-OFDM/SDMA is that its performance degrades when at least two users are received with approximately equal power. Under such transmission conditions, the probability of making an erroneous decision is increased, resulting in error propagation. This potential deficiency is illustrated in Figure 5 which shows for each carrier -in the bottom part-the signal-to-interference ratio (SIR) during the first iteration and -in the upmost part-the number of errors that occurred. Note however that from an information-theoretic viewpoint it is optimal that all users have identical power. To resolve the problem of error propagation, we use interference-dependent state insertion (SI). Essentially, by inserting additional state information on those carriers that suffer from a bad SIR, we can seriously decrease the probability of error propagation at a reasonable cost. Preferably State Insertion is applied in the (pc)SIC-OFDM/SDMA methods. The State Insertion method is further presented in a subcarrier by subcarrier approach but is not limited hereto.

[0093] First, the SIRs of user 1 for each subcarrier n are calculated from the knowledge of the equalizers $E_{LS}[s]$. Next, one additional state m is assigned to each of the M subcarriers n_m that suffer most from SIR. These M states keep track of alternative estimates, called intermediate hard estimates in the above, $\bar{x}_M^1[m]$ for user 1. These are defined as the nearest constellation point to $\tilde{x}_M^1[s_m]$ except for $\bar{x}^1[s_m]$, which is the sliced version of $\tilde{x}^1[s_m]$. After these assignments the M additional states are treated as normal subcarriers to obtain $\bar{x}_M^u[s_m]$, successively for u=2 to U. In explicit, the MUI originating from user u-1 is reconstructed (after recombining with a channel response estimation) and subtracted (thereby modifying the original signal) according to Formula 12 and the soft estimates $\tilde{x}_M^u[s_m]$ are computed by least-squares combining according to Formula 13. Finally, for each of those M subbands the algorithm selects from the intermediate hard estimates either $\bar{x}^u[s_m]$ or $\bar{x}_M^u[s_m]$ as hard estimate if respectively the first or second of the 2-norms of Formula 13 is smallest.

$$\begin{bmatrix} y_{1_M}^{u}[m] \\ \vdots \\ y_{A_M}^{u}[m] \end{bmatrix} = \begin{bmatrix} y_{1_M}^{u-1}[m] \\ \vdots \\ y_{A_M}^{u-1}[m] \end{bmatrix} - \begin{bmatrix} h_{1_M}^{u-1}[s_m] \\ \vdots \\ h_{A_M}^{u-1}[s_m] \end{bmatrix} \bar{x}_M^{u-1}[m] \qquad \text{Formula 12}$$

$$\tilde{x}_M^u[m] = \left[e_1^u[s_m]\cdots e_A^u[s_m]\right]\cdot\begin{bmatrix} y_{1,M}^{u}[m] \\ \vdots \\ y_{A,M}^{u}[m] \end{bmatrix} \qquad \text{Formula 13}$$

[0094] The State Insertion uplink transmission method can be described as a method wherein said determining of said estimates of said data signals from said subband processed received data signals in said receiving terminal further comprising for at least one data signal (but not limited to one) the following steps: (first step) selecting from said data signals a selected data signal; (second step) determining a plurality of estimates of said selected data signal from said subband processed received data signals, said estimates are above denoted intermediate hard estimates; (third step) determining a plurality of modified subband processed received data signals, each of said modified subband processed received data signals being based on one of said (intermediate hard) estimates of said selected data signal. A recombination and modifier method are exploited here; (fourth step) determining a plurality of estimates of at least one of said remaining data signals from each of said plurality of modified subband processed received data signals, for instance via but not limited to least squares linear combination; (last step) and thereafter selecting one of said estimates of said selected data signal, for instance but not limited to evaluation of a norm of said modified subband processed received data signals.

[0095] In an embodiment of State Insertion uplink transmission method said determining of estimates (all steps) is performed on a subband by subband basis.

[0096] In an embodiment of the invention said State Insertion uplink transmission method is used in combination with Successive Interference Cancellation, preferably on a subband by subband basis.

[0097] Figure 6 shows the performance of pcSIC-OFDM/SDMA with interference-dependent state insertion for the exemplary system. The number of extra states was set to M=64. As a reference, the dashed curve gives the performance of single-user single-antenna 100Mbps plain OFDM and the thin curves give the performance of pcSIC-OFDM/SDMA without state insertion. It shows that pcSIC-OFDM/SDMA with 64 additional states achieves a 6 dB gain at a BER of 0.001 and for four simultaneous users compared to pcSIC-OFDM/SDMA without SI. Other simulations also show that the performance improvement for SI with 32 and 16 states is respectively 5 dB and 4 dB.

[0098] For an estimation of the complexity of pcSIC-OFDM/SDMA with SI, the functionality can again be split in an initialization and a processing part. During initialization, each subcarrier's SIR after equalization needs to be computed, followed by state insertion for the M worst SIR subcarriers. During processing, M extra states need to be tracked using Formula 12 and 13, followed by selection based on the residual 2-norms of Formula 14.

$$\left\| \begin{matrix} y_1^{U+1}[s_m] \\ \vdots \\ y_A^{U+1}[s_m] \end{matrix} \right\|_2 \quad or \quad \left\| \begin{matrix} y_{1_M}^{U+1}[m] \\ \vdots \\ y_{A_M}^{U+1}[m] \end{matrix} \right\|_2 \qquad \text{Formula 14}$$

[0099] The operations required for state insertion of M out of N subcarriers are given in Figure 11. To obtain the total number of operations for pcSIC-OFDM/SDMA with SI the operations from Figure 10 have to be added to this. For the four-user four-antenna system, pcSIC-OFDM/SDMA with 64 additional states would require 220kflops and 310kdata-transfers during the initialization phase and 1.1Gflops/sec and a data-transfer bandwidth of 2.6Gwords/sec in the processing phase.

[0100] In the downlink transmission, the terminals of the composite peer are receiving signals and the processing peer is transmitting signals. The receiving terminals only dispose of a single antenna and a plain OFDM modem (i.e. without SDMA processing capabilities) in this embodiment. Since the user terminals have only a single antenna and since we want to concentrate most of the processing power in the basestation, they cannot mitigate multi-user interference in the downlink. Therefore, the basestation carries out a precompensation matrix E_{PRE} on the datasymbol vector Y[s] to obtain Y_{TR}[s], which is then transmitted on the A antennas. The input-output relation of Formula 6 is obtained.

[0101] Neglecting any non-idealities, this approach results in perfect separation of each user's datasymbols and perfect equalization upon reception. Therefore, no channel information or equalizer is needed in the mobiles. By applying these precompensation matrices for each carrier n, the input-output relation becomes as in Formula 15.

$$\begin{aligned} X[s] &= H^T[s] \cdot E_{PRE}[s] \cdot Y[s] + N''[s] \\ &= I_{U \times U} \cdot Y[s] + N''[s] \end{aligned} \qquad \text{Formula 15}$$

[0102] Thus, each user sees a single-user AWGN channel per carrier. Figure 7 shows the average BER in the downlink for different numbers of users. Since there is no interference, the number of simultaneous users does not influence performance (provided it does not exceed the number of antennas and no non-idealities occur).

[0103] The processing in the downlink again consists of both initialization and processing. In the initialization phase, the pre-compensation matrices are calculated. During the processing, the user data Y[s] is then multiplied with said pre-compensation matrices. The complexity for both initialization and processing is approximately the same as for the LS-OFDM/SDMA scheme in the uplink.

[0104] In addition to the proposed OFDM/SDMA techniques proposed for both uplink and downlink communication, the initialization effort in both situations can be simplified by applying coherence grouping. This initialization effort is characterized as determining the equalizer or the precompensation matrices, wherein Formulas 4, 5 or 7 are exploited. With said equalizer matrices relations between said data signals and subband processed received data signals are defined like in Formula 1, 9 or 13. With said precompensation matrices relations between said combined data signals, being transformed versions of said data signals, are defined like in Formula 6 or 15. Instead of determining said matrices for each subband separately, said subbands can be grouped into sets, at least one set comprising of at least two subbands and said matrices or more generally said relations can then be determined on a set-by-set basis. The groups are in an embodiment of this coherence grouping principle all of equal size G. Figure 8 shows the performance degradation associated with carrier grouping applied to pcSIC-OFDM/SDMA with SI, for several values of G. It is observed that for the simulated channel the degradation is negligible for G smaller then or equal to 4. For G greater then or equal to 32, the performance becomes inferior to single-user single-antenna plain OFDM. We may conclude that for the system of interest the initialization complexity can be reduced with a factor 8. For the example of a 100 Mbps OFDM/SDMA WLAN that uses SDMA to separate four simultaneous 25 Mbps users, some numerical results for the performance gain and the complexity can be given. If pcSIC-OFDM/SDMA with 64-SI and a group size of eight is implemented, a performance gain of 11dB is obtained at a BER of 0.001 compared to the least-squares approach. The required computational power is respectively 27kflops and 40kdata-transfers during initialization and 1.1Gflops/sec and a data-transfer bandwidth of 2.6Gwords/sec during processing.

[0105] In another preferred embodiment, the processing peer provides an access point to a connecting network (this connecting network can for example be a cable network or a satellite network), and the composite peer consists of wireless terminals for which cost is a major issue. In this embodiment, only the access point has DFT (Discrete Fourier Transform) and IDFT (Inverse Discrete Fourier Transform) processing means. This enables a cheap realization of both the digital baseband modem part of the terminals in the composite peer, and of the front-ends of all terminals in the system since a high peak-to-average power ratio of the transmitted signals is avoided. A typical application of asymmetric frequency domain SDMA could for example be a wireless home networking application in the 2.4 GHz band. Such an asymmetric configuration does implements a so-called concentrated scenario, wherein subband processing and inverse subband processing is located in the processing peer. In this transmission method the determination of estimates of said data signals from said subband processed received data signals in said receiving terminal comprising the following steps: (first step) determining intermediate estimates of said data signals from said subband processed received data signals in said receiving terminal; (second step) obtaining said estimates of said data signals by inverse subband processing said intermediate estimates. In this embodiment, the methods of the invention can be used not only to detect the signals of different terminals transmitting simultaneously in the same frequency band, but for example also to mitigate the interference from a microwave oven causing disturbance signals in the band. In multipath fading environment, such as for example home environments, guard intervals can be inserted in the data signals prior to transmission. If these guard intervals are again designed to comprise all the echoes received on the multipath channel, the channel convolution becomes cyclic after removal of the guard intervals. Thus, in the frequency-domain it becomes equivalent to multiplication with the Fourier transform of the channel. This embodiment allows cheap terminals in the processing peer and can implement uplink as well as downlink combinatory processing, in a way similar as OFDM/SDMA, including successive interference cancellation, preferably per subband and state insertion.

Claims

1. A method of transmitting data signals (50) from at least two transmitting terminals (20) with each at least one transmitting means (60) to at least one receiving terminal (40) with a spatial diversity receiving means (80) comprising the steps:

   - transmitting from said transmitting terminals (20) transformed data signals (70), being transformed versions of said data signals; receiving on said spatial diversity means (80) received data signals being at least function of at least two of said transformed data signals (70);
   - subband processing (90) of at least two of said received data signals in said receiving terminal (40); and

- determining estimates of said data signals (120) from subband processed received data signals (140) in said receiving terminal.

2. The method recited in 1, wherein said transmissing being substantially simultaneously.

3. The method recited in 1, wherein, the spectra of said transformed data signals being at least partly overlapping.

4. The method recited in 1, wherein the spectra of said transformed data signals being at least partly overlapping.

5. The method recited in 1, wherein the step of determining of said estimates of said data signals from subband processed received data signals in said receiving terminal further comprising for at least one data signal the steps of:

   - selecting from said data signals a selected data signal;
   - determining an estimate of said selected data signal from said subband processed received data signals;
   - modifying said subband processed received data signals based on said estimate of said selected data signal; and
   - determining estimates of said remaining data signals from said modified subband processed received data signals.

6. The method recited in claim 5, wherein the step of selecting a data signal being based on the receiving power of said data signals.

7. The method recited in claim 5, wherein the step of selecting a data signal being based on the interference ratio of said data signals.

8. The method recited in claim 1, wherein the step of determining of said estimates of said data signals from subband processed received data signals in said receiving terminal further comprises for at least one data signal the steps of:

   - selecting from said data signals a selected data signal;
   - determining a plurality of estimates of said selected data signal from said subband processed received data signals;
   - determining a plurality of modified subband processed received data signals, each of said modified subband processed received data signals being based on one of said estimates of said selected data signal;
   - determining a plurality of estimates of at least one of said remaining data signals from said plurality of modified subband processed received data signals; and
   - thereafter selecting one of said estimates of said selected data signal.

9. The method recited in claim 8, wherein the step of selecting a data signal being based on the interference ratio of said data signals.

10. The method recited in claim 1, wherein:

   - the subbands, being involved in said subband processing, being grouped into sets, at least one set comprising of at least two subbands;
   - the step of determining of said estimates of said data signals in said receiving terminal comprising the steps:
   - determining relations between said data signals and subband processed received data signals on a set-by-set basis;
   - exploiting said relations between said data signals and said subband processed received data signals for determining said data signals.

11. The method recited in claim 1, wherein said transformation of said data signals (50) to transformed data signals (70) comprising the step of inverse subband processing (160).

12. The method recited in claim 1, wherein the step of determining (150) estimates of said data signals from subband processed received data signals in said receiving terminal comprises the steps of:

   - determining (100) intermediate estimates of said data signals (130) from said subband processed received

data signals in said receiving terminal;

- obtaining said estimates of said data signals (120) by inverse subband processing (110) said intermediate estimates.

13. The method recited in claim 1, wherein said transformation of said data signals to transmitted data signals further comprising guard interval introduction.

14. The method recited in claim 1, wherein said subband processing being orthogonal frequency division demultiplexing.

15. The method recited in claim 11 and 12, wherein said inverse subband processing being orthogonal frequency division multiplexing.

16. A method of transmitting data signals (200) from at least one transmitting terminal (240) with a spatial diversity transmitting means (220) to at least two receiving terminals (330)with at least one receiving means (320) comprising the steps:

- determining (250) combined data signals (300) in said transmitting terminal, said combined data signals being transformed versions of said data signals;
- inverse subband processing (260) said combined data signals;
- transmitting with said spatial diversity means (220) inverse subband processed combined data signals;
- receiving on at least one of said receiving means (320) of at least one of said receiving terminals (330) inverse subband processed received data signals, being at least function of said inverse subband processed combined data signals; and
- determining estimates of said data signals from said inverse subband processed received data signals.

17. The method recited in claim 16, wherein said transmitting of inverse subband processed combined data signals being substantially simultaneously.

18. The method recited in claim 16, wherein the spectra of said inverse subband processed combined data signals being at least partly overlapping.

19. The method recited in claim 16, wherein the step of determining of combined data signals in said transmitting terminal being on a subband by subband basis.

20. The method recited in claim 16, wherein the step of determining of said estimates of said data signals in said receiving terminals comprising subband processsing (350).

21. The method recited in claim 16, wherein the step of determining combined data signals in said transmitting terminal comprising the steps:

- determining intermediate combined data signals (290) by subband processing (280) said data signals;
- determining (270) said combined data signals from said intermediate combined data signals.

22. The method recited in claim 20 and 21, wherein said subband processing being orthogonal frequency division demultiplexing.

23. The method recited in claim 16, wherein said inverse subband processing being orthogonal frequency division multiplexing.

24. The method recited in claim 16, wherein:

- said subbands, being involved in inverse subband processing, being grouped into sets, at least one set comprising of at least two subbands;
- the step of determining (250) of combined data signals (300) in said transmitting terminal (240) comprising the steps:
- determining relations between said data signals and said combined data signals on a set-by-set basis; and
- exploiting said relations between said data signals and said combined data signals for determining said data

signals.

25. The method recited in claim 16, wherein in said inverse subband processed combined data signals a guard interval being introduced.

26. An apparatus for determining estimates of data signals from at least two received data signals, said received data signals, said apparatus comprising at least of

- at least one spatial diversity receiving means (80);
- circuitry being adapted for receiving said received data signals with said spatial diversity receiving means;
- circuitry being adapted for subband processing (90) at least two of said received data signals; and
- circuitry (150) being adapted for determining estimates of said data signals from subband processed received data signals.

27. The apparatus recited in claim 26 wherein said circuitry being adapted for determining estimates of said data signals from subband processed received data signals comprises a plurality of circuits each being adapted for determining part of said estimates of said data signals based on part of the subbands of said subband processed received data signals.

28. The apparatus recited in claim 26, wherein said spatial diversity means comprises of at least two receiving means and said circuitry being adapted for receiving said received data signals with said spatial diversity means comprises a plurality of circuits each being adapted for receiving said received data signals from one of said receiving means of said spatial diversity means.

29. An apparatus for transmitting inverse subband processed combined data signals comprising at least of:

- at least one spatial diversity transmitting means;
- circuitry being adapted for combining data signals;
- circuitry being adapted for inverse subband processing combined data signals;
- circuitry being adapted for transmitting inverse subband processed combined data signals with said spatial diversity means.

30. The apparatus recited in claim 29, wherein said circuitry being adapted for combining data signals comprising a plurality of circuits each being adapted for combining data signals based on part of the subbands of said data signals.

31. The apparatus recited in claim 29, wherein said spatial diversity transmitting means comprises at least two transmitting means and said circuitry being adapted for transmitting inverse subband processed combined data signals comprises a plurality of circuits each being adapted for transmitting said inverse subband processed combined data signals with one of said transmitting means of said spatial diversity means.

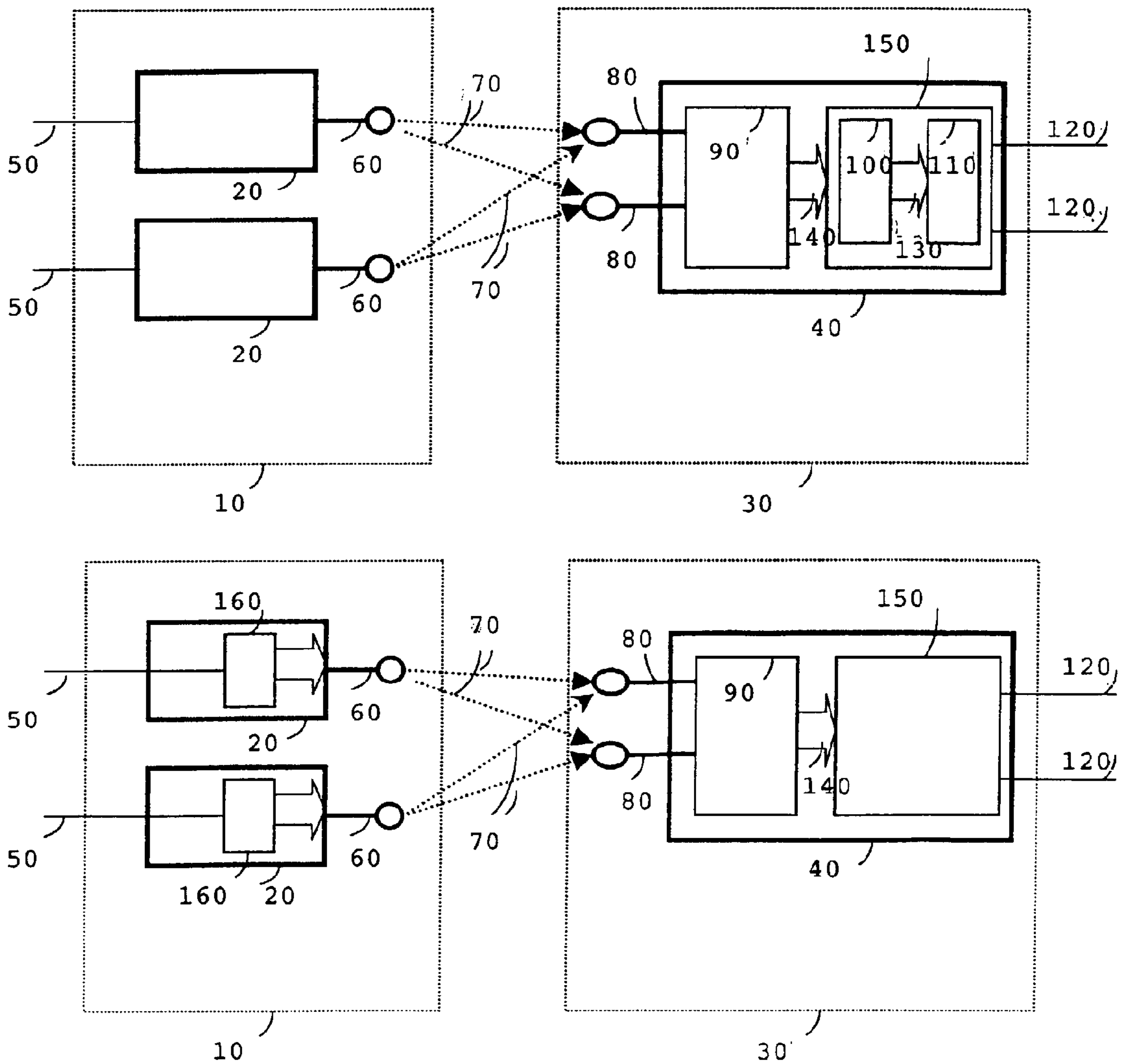
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FIG. 1

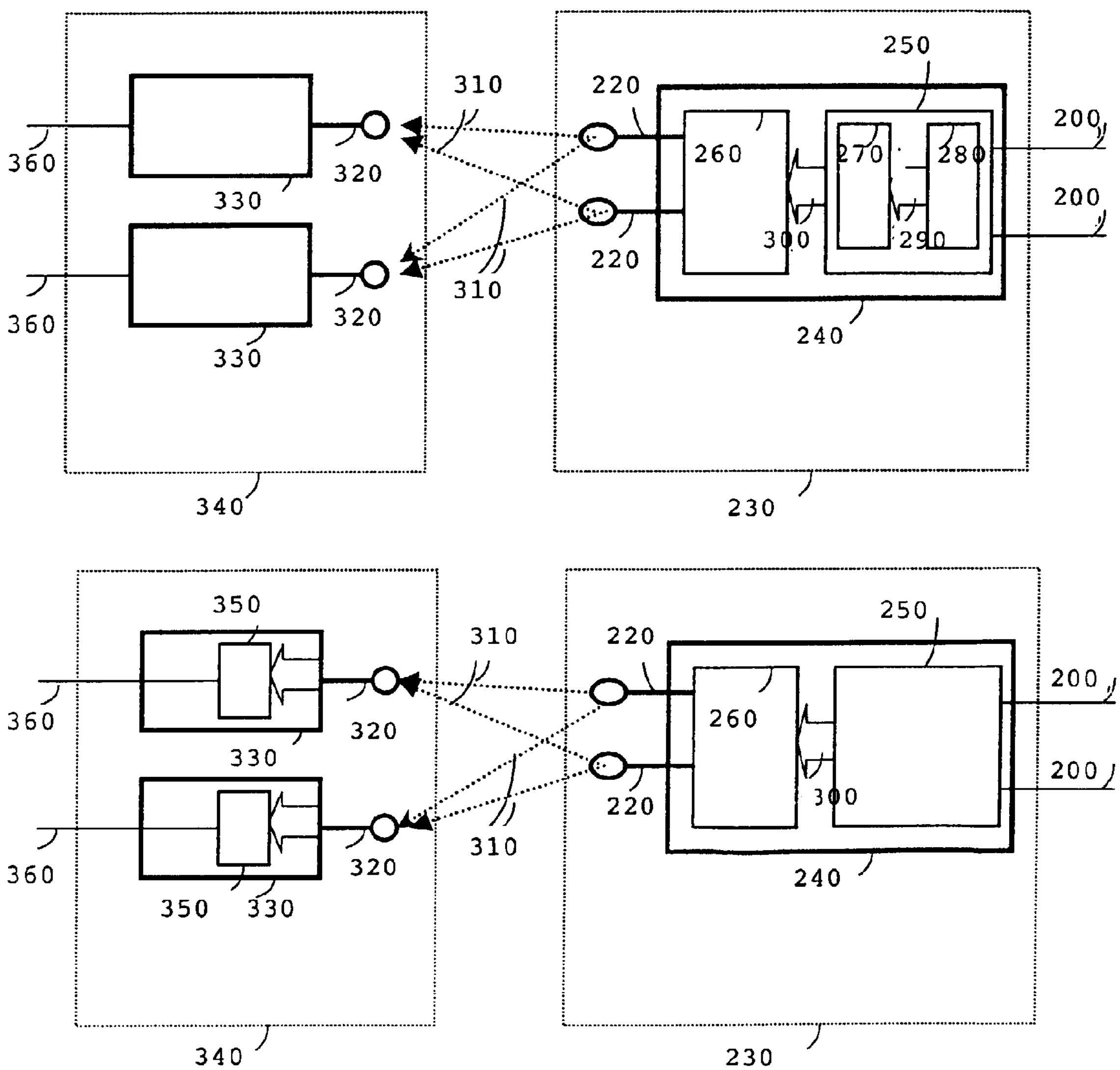
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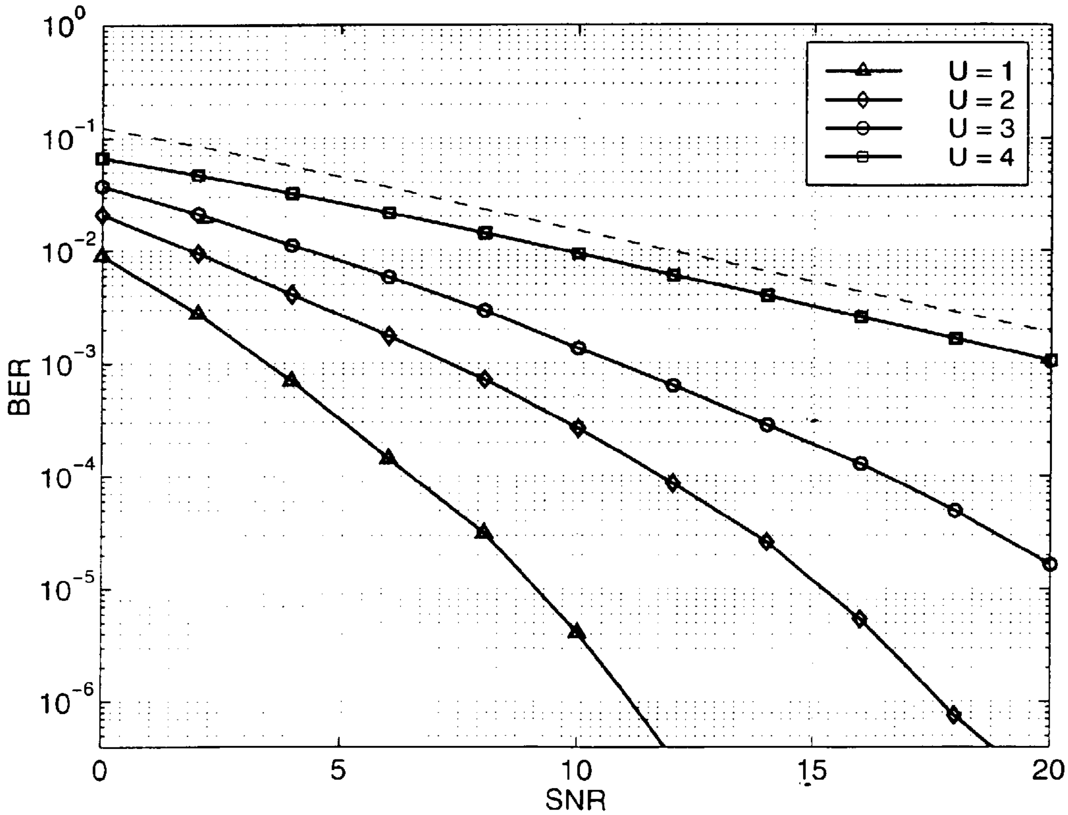
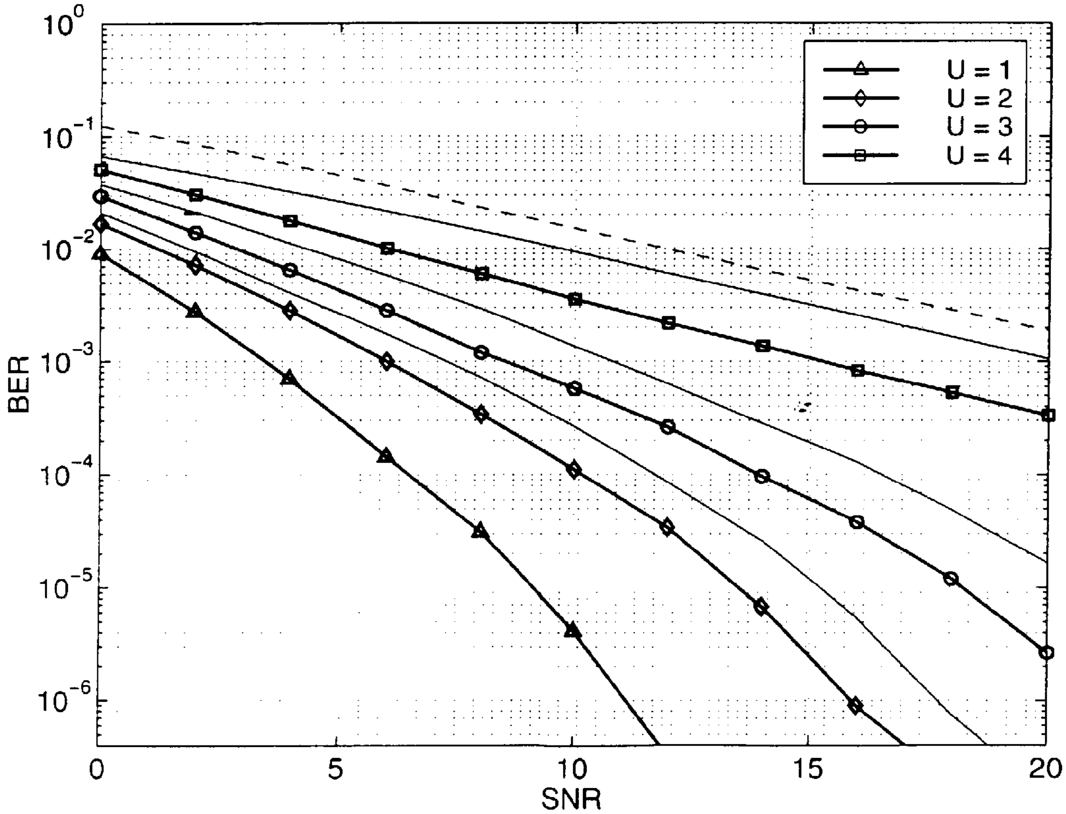
FIG. 2



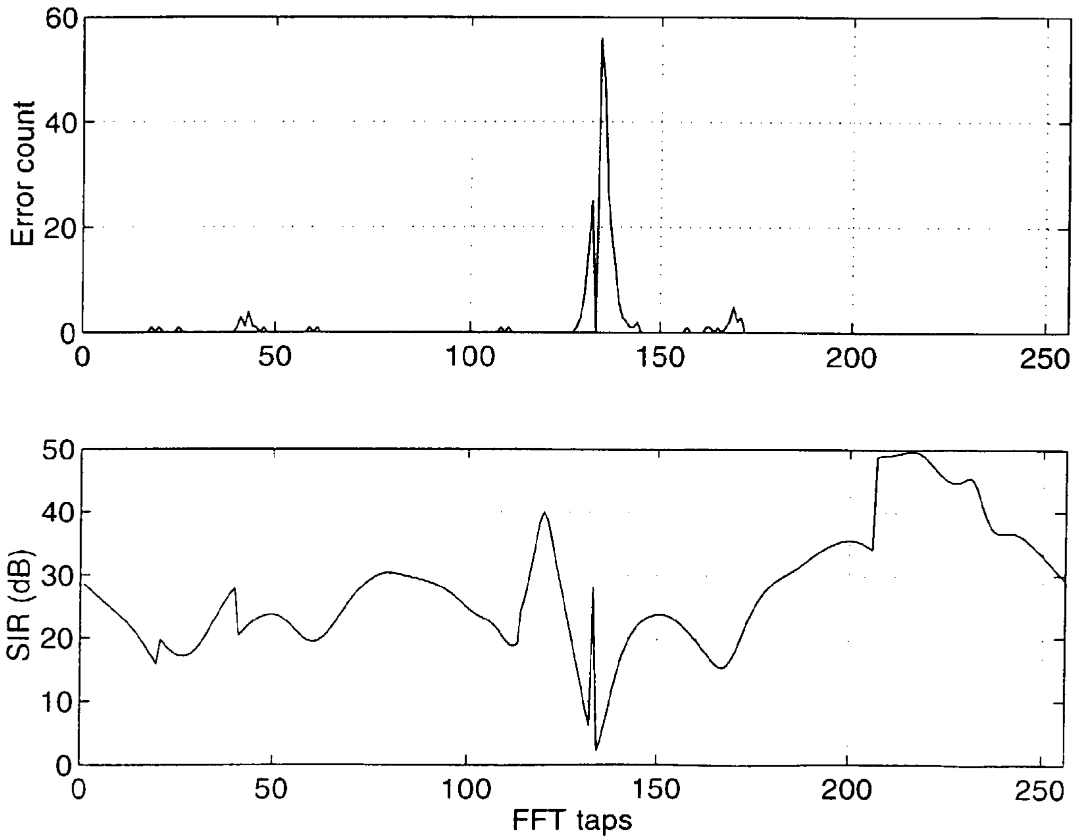
Performance of LS-OFDM/SDMA

FIG. 3



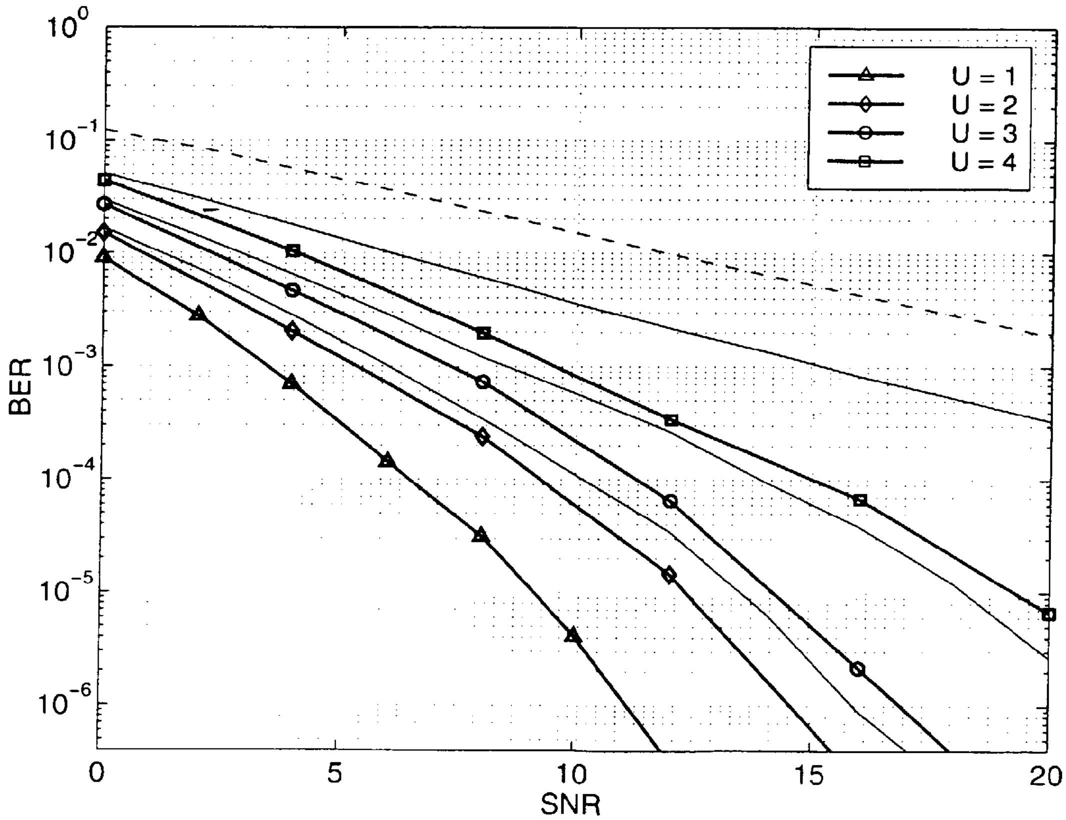
Performance of pcSIC for OFDM/SDMA

FIG. 4



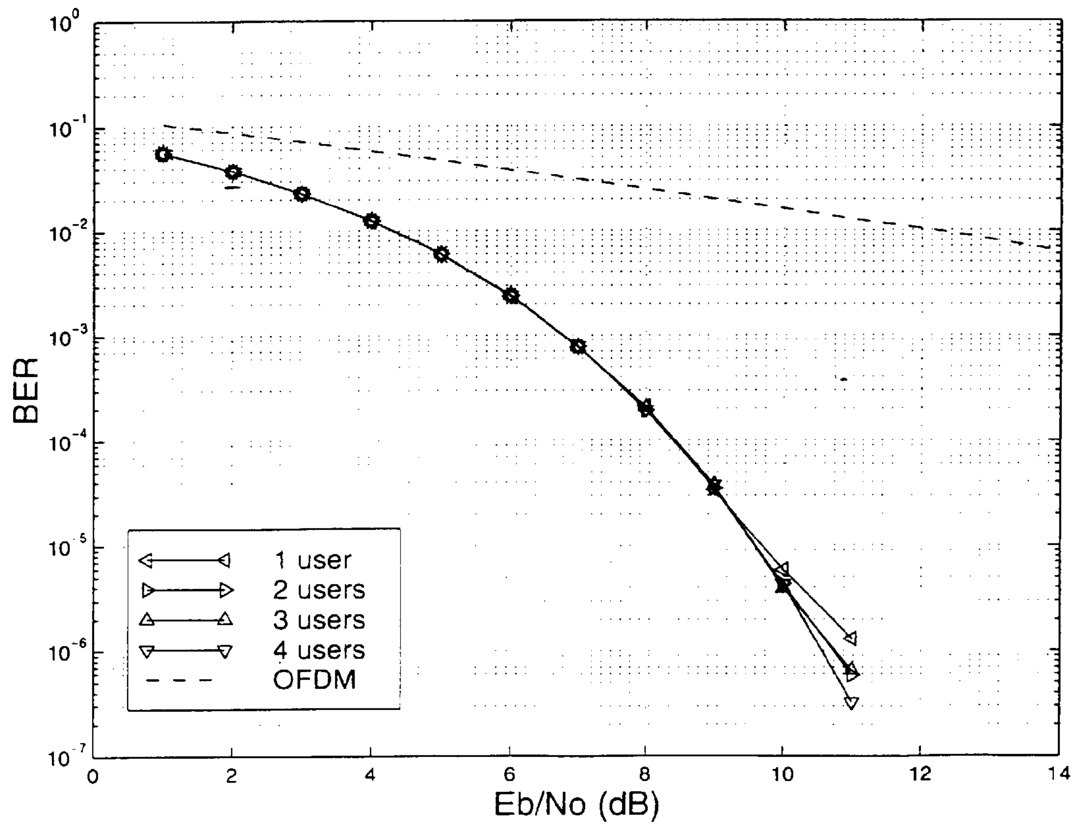
Error propagation with SIC-OFDM/SDMA

FIG. 5



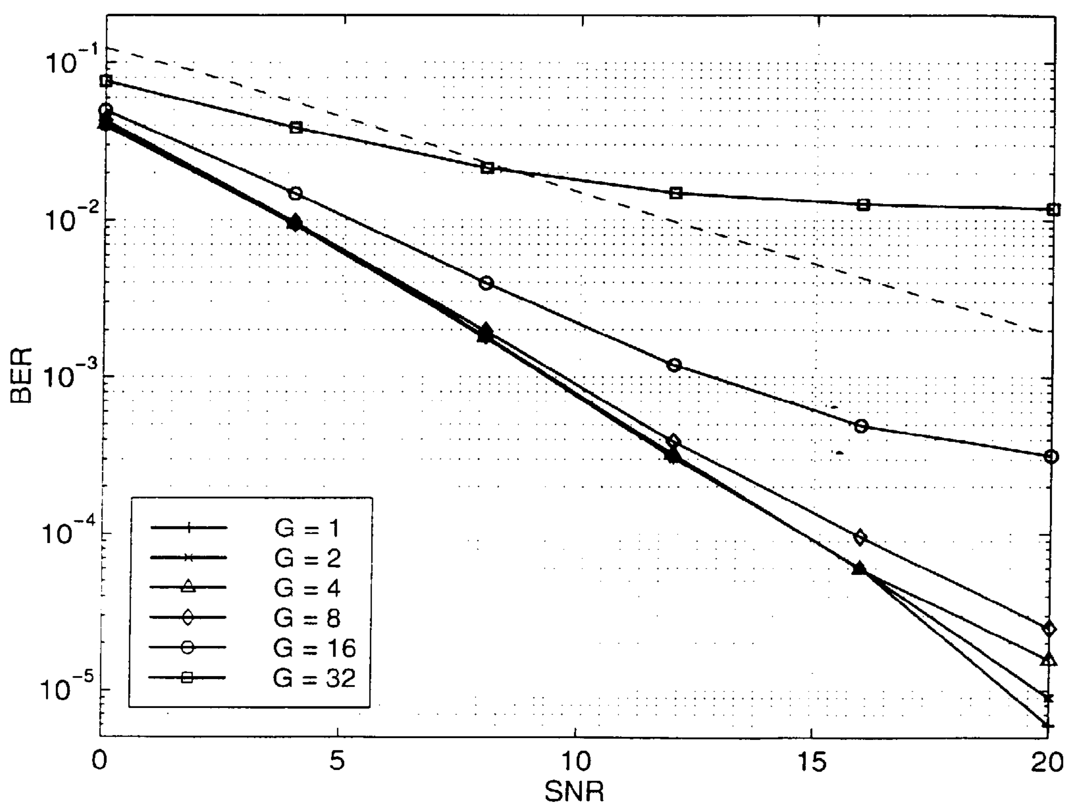
Performance of pcSIC-OFDM/SDMA with SI

FIG. 6



Performance of OFDM/SDMA in the downlink

FIG. 7



Coherency Grouping

FIG. 8

LS-OFDM/SDMA operation count

	initialization	processing
multiplications	$A^3/3 + 2A^2U$	AU
additions	$A^3/3 + 2A^2U$	AU
data transfers	$A^3 + 9A^2U$	$6AU$

FIG. 9

pcSIC-OFDM/SDMA operation count

	initialization	processing
multiplications	$A^3U/3 + 12A^2U$	$2AU - A$
additions	$A^3U/3 + 12A^2U$	$2AU - A$
data transfers	$A^3U/3 + 24A^2U$	$10AU - 4A$

FIG. 10

additional operation count for SI

	initialization	processing
multiplications	$AU^2/2$	$2(M/N\,AU + A)$
additions	$(AU^2 + N)/2$	$2(M/N\,AU + A)$
data transfers	$AU^2 + N/2$	$10(M/N\,AU + A)$

FIG. 11

European Patent Office

EUROPEAN SEARCH REPORT

Application Number
EP 99 87 0168

Category	DOCUMENTS CONSIDERED TO BE RELEVANT Citation of document with indication, where appropriate, of relevant passages	Relevant to claim	CLASSIFICATION OF THE APPLICATION (Int.Cl.7)
X Y	WO 97 41647 A (BAIER PAUL WALTER ;KAMMERLANDER KARL (DE); JUNG PETER (DE); SIEMEN) 6 November 1997 (1997-11-06) * abstract * * page 2, line 27 - page 7, line 22 * * figure 1 * ---	1-3,11, 13-15,26 5,8,10, 12,27,28	H04B7/08 H04L27/26 H04L1/06
Y	FAZEL K. ; FETTWEIS G.P.: "Multi-Carrier Spread Spectrum" 1997 , KLUWER ACADEMIC PUBLISHERS , DORDRECHT, THE NETHERLANDS XP002130135 * page 50, paragraph 2 - page 52, paragraph 2 * * figures 1,2 * ---	16-25, 29-31	
Y Y	KUZMINSKIY, A.M.; HATZINAKOS, D.: "Multistage semi-blind spatio-temporal processing for short burst multiuser SDMA systems" CONFERENCE ON SIGNALS, SYSTEMS & COMPUTERS, vol. 2, 1 - 4 November 1998, pages 1887-1891, XP002130134 * abstract * * page 1888, column 1, paragraph 1 - page 1890, column 1, paragraph 1 * ---	5,8,10, 12,27,28 16-25, 29-31	TECHNICAL FIELDS SEARCHED (Int.Cl.7) H04B H04L
A	WO 98 51048 A (NEILL RORIE JULIAN TURNBULL O ;MOTOROLA LTD (GB)) 12 November 1998 (1998-11-12) * page 3, line 14 - page 4, line 27 * * page 7, line 14 - page 8, line 30 * * page 10, line 1-21 * * figures 3-5 * -----	1-3,11, 13-15,26	

The present search report has been drawn up for all claims

Place of search	Date of completion of the search	Examiner
THE HAGUE	9 February 2000	Yang, Y

CATEGORY OF CITED DOCUMENTS

X : particularly relevant if taken alone
Y : particularly relevant if combined with another document of the same category
A : technological background
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P : intermediate document

T : theory or principle underlying the invention
E : earlier patent document, but published on, or after the filing date
D : document cited in the application
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ANNEX TO THE EUROPEAN SEARCH REPORT ON EUROPEAN PATENT APPLICATION NO. EP 99 87 0168

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09-02-2000

Patent document cited in search report	Publication date	Patent family member(s)	Publication date
WO 9741647 A	06-11-1997	DE 19616829 C	24-04-1997
		EP 0895683 A	10-02-1999
WO 9851048 A	12-11-1998	GB 2324932 A	04-11-1998
		AU 7648998 A	27-11-1998
		EP 0922350 A	16-06-1999

EPO FORM P0459

For more details about this annex : see Official Journal of the European Patent Office, No. 12/82

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(11) EP 1 087 545 A1

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(71) Applicant:
NATIONAL UNIVERSITY OF SINGAPORE
Singapore 119260 (SG)

(72) Inventors:
- **Liang, Ying Chang**
  10-559 Singapore 120301 (SG)
- **Chin, Francois Po Shin**
  02-228 Singapore 550208 (SG)

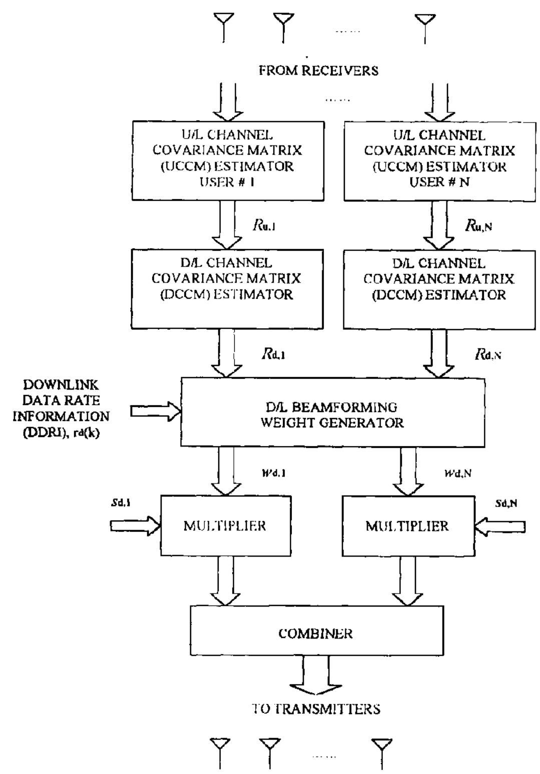
(74) Representative:
Hoarton, Lloyd Douglas Charles
Forrester & Boehmert,
Franz-Joseph-Strasse 38
80801 München (DE)

(54) Downlink beamforming method

(57) A method for downlink capacity enhancement in a wireless communications system comprising a base station with antenna array and terminals that are physically remote from said base station, the method comprising steps of:

receiving at said base station antenna array combinations of arriving signals from said plurality of remote terminals;
estimating an uplink channel covariance matrix (UCCM) for each of said terminals from said combinations of arriving signals;
constructing from each of said UCCM a downlink channel covariance matrix (DCCM);
calculating from all said DCCM a downlink weight vector for each of said terminals; and
transmitting a set of information signals from said base station antenna array according to said downlink weight vectors.

FIG. 2



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Description

BACKGROUND OF THE INVENTION

[0001] The present invention relates in general to wireless communication systems and, more particularly, to using antenna array and signal processing techniques to increase downlink capacity and performance of wireless communication systems.

[0002] The next generation of wireless mobile communication systems will be required to provide a variety of services including digital voice, video and data in different transmission modes. These services will require higher data rates and higher received signal power levels, thus creating increased interference between users. In order to obtain high system capacity, the interference levels have to be reduced dramatically. Spatial division multiple access (SDMA), by which a plurality of antenna elements are equipped at the base station in order to receive and transmit data information from and to the desired user by using spatial diversities, has been proposed as an effective technique to achieve this.

[0003] The main operations in SDMA include uplink (from mobile station to base station) beamforming and downlink (from base station to mobile station) beamforming. Uplink beamforming consists of uplink beamforming weight generation and uplink signal demultiplexing; Downlink beamforming includes downlink beamforming weight generation and downlink signal multiplexing. Theoretically, in both links, the associated channel responses are of critical importance in order to generate corresponding beamforming weights.

[0004] Usually the antenna array is equipped at base station, not at mobile terminals due to size limitation. Uplink beamforming is easy for implementation since uplink channel responses (UCRs) can be directly measured. Therefore much attention has been paid to uplink capacity enhancement. However, it is also desirable to improve downlink capacity in order to improve the whole system capacity. Moreover, downlink capacity is even more important for the next generation mobile communication systems in which wireless internet, video-on-demand and multimedia services are to be required.

[0005] In wireless communications, two duplex modes can be used: time-division-duplex (TDD) and frequency-division-duplex (FDD). For TDD mode, uplink and downlink channel responses are equal if the dwelling time is short enough. Thus UCRs can be used as downlink channel responses (DCRs) in determining downlink beamforming weights. This approach, however, requires accurate synchronization between uplink and downlink time slots, otherwise, interference between uplink and downlink signals can be seriously large. For FDD mode, since uplink and downlink employ different carrier frequncies, uplink and downlink signals will not interfere with each other. Therefore, FDD duplex mode is adopt in most current wireless communication systems, and most probably will be used in the next generation systems.

[0006] In FDD systems, UCRs are different from DCRs since the RF propagation environment differs at the uplink and downlink carrier frequencies. Hence, using antenna array at the base station to improve downlink performance is usually a more difficult problem than the associated uplink one due to lack of direct measurement of downlink channel responses (DCRs). In U.S. Patent No. 5,471,647, D. Gerlach and A. Paulraj proposed one conceptually simple method, called probing-feedback approach, to estimate DCRs. In this approach, probing signals are first sent to the mobile users from base station in order to measure the instantaneous downlink channel vectors (IDCVs), then the IDCVs are feedback to the base station to generate downlink beamforming weights using certain criterion. This approach, however, is only applicable in environment which varys very slow in time. In another U.S. Patent No 5,634,199, D. Gerlach and A. Paulraj proposed to feedback the stable downlink channel vectors (SDCVs) in order to reduce the feedback rate. Both methods seem to be not advisable since they require complete redesign of uplink and downlink protocols and signaling. Moreover, these methods may greatly reduce the transmission and spectrum efficiency.

[0007] Another kind of approach for estimating DCRs is based upon direction-of-arrival (DOA) information embedded in received uplink signals. In fact, since uplink and downlink signals travel through reflections and deflections due to same scatters surrounding the mobile and the base station, the DOAs of the uplink signals might be the only constant parameters which can be used for downlink beamforming.

[0008] DOA-based approaches employ the received uplink signals to compute the desired user's DOAs first; then DCRs are estimated by constructing downlink steering vectors for given DOAs. In International Patent Application Publication No. WO 97/45968, "Method of and apparatus for interference rejection combining and downlink beamforming in a cellular radiocommunications system", (12/97), Forssen *et al* proposed to compute the probability function with respect to different DOAs at which the desired signal may come from, and to choose the angle of incidence associated with the particular mobile station as the DOA value which maximizes the probability function. This technique, however, suffers from heavy computational burden in computing the probability function and searching the maximum point. In another International Patent Application Publication No. WO 96/22662, "Spectrally efficient high capacity wireless communication systems", (7/96), Barratt *et al* use subspace-based techniques (*e.g.*, MUSIC and ESPRIT) to obtain high-resolution DOA estimates from the covariance matrix of the antenna outputs. It is well known that subspace-based algorithms require very complicated computations since they are involved in the computation of matrix inversion or singular

value decomposition of complex matrices, and one or even more multidimensional nonlinear optimizations. On the other hand, accurate DOA estimates are not available in multipath cases since the number of multipath DOAs are usually greater than the number of antenna elements. This may limit the applicability of the DOA-based approaches for estimating DCRs.

[0009] In fact, from U.S. Patent No. 5,634,199, it is the downlink channel covariance matrices (DCCMs) that determine the downlink beamforming weights. Similar conclusions were drawn and exploited by C. Farsakh and J.A. Nossek in paper, "Spatial covariance based downlink beamforming in an SDMA mobile radio system", *IEEE Trans. Comms.*, vol.46, No. 11, 1998, pp.1497-1506. However, besides probing-feedback approach, the above two literatures failed to provide any efficient technique to compute DCCMs for FDD systems. Although in paper, "Downlink beamforming for spatially distributed sources in cellular mobile communications", *Signal Processing*, Vol.65, 1998, 181-197, Goldberg and Fonollosa proposed a method for estimating DCCMs. This technique, however, also suffers from heavy computational burden and there is room to further simplify the computation of DCCM so that it is easier for practical implementation. Yet, the approach proposed by Goldberg and Fonollosa cannot be applied to the cases in which receive and transmit antenna structures are different from each other.

[0010] As such, the first objective of the present invention is to develop a computationally efficient technique for generating DCCMs and SDCVs for FDD systems.

[0011] Once DCCMs or SDCVs are obtained, the work left is to design downlink beamforming weights using DCCMs or SDCVs. Traditional approach is to use SDCVs as the downlink weight vectors. This approach, called maximal ratio combining (MRC) approach, is equivalent to keeping the main beam of the downlink beam pattern toward the intended user. Since uplink usually employs minimum mean-square-error (MMSE) beamforming scheme, which is much better than MRC method, the traditional approach is not able to provide enough capacity to match its uplink counterpart. Another approach is proposed by F. Rashid-Farroki *et al* in paper, "Transmit beamforming and power control for cellular wireless systems," *IEEE Journal of Selected Areas in Communications*, vol. 16, No.8, Oct. 1998, pp.1437-1449. This approach generates downlink beamforming weights using joint uplink beamforming and power control technique in which total transmitted power is to be minimized. This approach, however, does not consider data rate information, and more seriously, no efficient technique is suggested to jointly solve FDD and weight generation problem.

[0012] The next generation systems will be required to provide wireless internet, video-on-demand and multimedia services, thus users sharing the same channel may request higher data rates and higher received signal powers. If each user's main beam is simply directed to the desired user without considering the interference polluted to the other users, the quality of the low rate user spatially closed to stronger users may be so poor that even the minimum quality requirement cannot be satisfied. Thus how to design downlink beamforming weights such that maximum number of users with different data rate services can be supported within the same channel and same cell or sector while keeping satisfactory communication quality becomes the second objective of the present invention.

[0013] As mentioned earlier, in SDMA wireless communications, the main operations include uplink weight generation and downlink weight generation. Since uplink beamforming weights are useful information at hand, the third objective of the present invention is to develop methods for generating downlink beamforming weights by direct modifying uplink ones.

SUMMARY OF THE INVENTION

[0014] The present invention comprises a wireless communication system which integrates base station antenna array and signal processing techniques to improve downlink performance and capacity of wireless communications.

[0015] According to the present invention, an apparatus for communicating with a plurality of wireless users is provided which consists of uplink receive antenna array, uplink weight generator and uplink spatial demultiplexing, and downlink weight generator, downlink spatial multiplexing and downlink transmit antenna array. Downlink beamforming weights can be derived from uplink channel covariance matrices (UCCMs), or uplink channel responses (UCRs), or uplink beamforming weights. Thus no feedback or intermediate step for estimating DOAs is required. Also, downlink transmit antenna array can be same as or different from uplink receive antenna array.

[0016] According to one aspect of the present invention, uplink receive antenna array acquires a plurality of combinations of signals transmitted from the mobile users, from which UCRs or UCCMs are estimated. Downlink channel covariance matrices (DCCMs) or downlink channel responses (DCRs) can then be derived from UCCM or UCRs together with certain frequency calibration processing.

[0017] Advantageously, DCCMs can be estimated from UCCMs via peak constraint method. Peak constraint method generates DCCM by keeping same peak positions of main beams for the beam patterns generated from the eigenvectors of UCCM and DCCM. This method links columnized DCCM vector with columnized UCCM vector through a linear multiplication with a frequency calibration (FC) matrix, which is only dependent on uplink and downlink carrier frequencies, receive and transmit antenna array structures of the system, and can be computed and stored in advance. Thus, peak constraint method is a simple while efficient technique for overcoming FDD problem.

[0018] Conveniently, SDCVs can be estimated from UCRs using peak constraint or null constraint methods. For peak constraint method, the principal eigenvector of the estimated DCCM is used as SDCV. For null constraint, SUCV is first estimated by computing the principal eigenvector of the UCCM, which is obtained from IUCVs via time-average approach, then SDCV is generated by keeping same null positions for the beam patterns generated from both SUCV and SDCV.

[0019] According to one aspect of the present invention, downlink beamforming weights can be generated from DCCMs or SDCVs using different approaches, such as iterative virtual power weighted (IVPW) approach, virtual power weighted (VPW) approach or spatial distribution weighted (SDW) approach. Downlink data rate information is exploited in designing downlink beamforming weights in order to maximize the system capacity.

[0020] According to one aspect of the present invention, downlink beamforming weights can be generated by direct modifying uplink beamforming weights.

[0021] Preferably, downlink beamforming weights can be implemented by using normal uplink beamforming weights together with null constraint method. The method is simple for implementation in terms of hardware and software complexities since uplink beamforming weights are already at hand.

[0022] Advantageously, downlink beamforming can be implemented by using leaky uplink beamforming weights together with certain frequency calibration processing, such as peak constraint transform. This method provides another choice for implementing downlink beamforming since in some cases leaky uplink beamforming scheme is already used in order to keep the uplink weight from converging to pathological solutions.

[0023] The basic properties and benefits of the present invention are summarized as follows:

1. The present invention provides a high flexibility in the sense that different kind of uplink information can be used, such as uplink channel covariance matrices, uplink channel responses and uplink beamforming weights.
2. The present invention is simple to implement. It does not require downlink channel feedback, thus eliminating the need for modifying uplink and downlink protocols, and not require demanding DOA estimation and its association.
3. The main concern complicating FDD system is the lack of downlink channel responses. The present invention provides two methods for solving this problem: peak constraint method and null constraint method.
4. The present invention takes care of possibly different receive and transmit antenna array structures, no matter if the systems are TDD or FDD.
5. The present invention provides different methods for generating downlink beamforming weights based upon different uplink information used. Downlink data rate information is also exploited in order to maximize system capacity. These methods can be applied in both TDD and FDD systems.

BRIEF DESCRIPTION OF DRAWINGS

[0024]

FIG. 1 is a graphic illustration of prior downlink beamforming scheme,
FIG.2 is a graphic illustration of the downlink beamforming scheme using uplink channel covariance matrix (UCCM) estimates in accordance with the embodiment 1 of the present invention;
FIG.3 is a block diagram of DCCM estimator in accordance with the present invention;
FIG.4 shows first embodiment of downlink beamforming generator using DCCMs;
FIG.5 shows second embodiment of downlink beamforming generator using DCCMs;
FIG.6 shows third embodiment of downlink beamforming generator using DCCMs;
FIG.7 shows fourth embodiment of downlink beamforming generator using DCCMs;
FIG.8 is a graphic illustration of the downlink beamforming scheme using uplink channel responses (UCRs) in accordance with embodiment 2 of the present invention;
FIG.9 shows one embodiment of stable downlink channel vector (SDCV) estimator,
FIG.10 shows another embodiment of SDCV estimator;
FIG.11 shows first embodiment of downlink beamforming generator using SDCVs;
FIG.12 shows second embodiment of downlink beamforming generator using SDCVs;
FIG.13 shows third embodiment downlink beamforming generator using SDCVs,
FIG.14 is a graphic illustration of the downlink beamforming scheme using normal uplink weights in accordance with embodiment 3 of the present invention;
FIG.15 illustrates one embodiment of downlink beamforming weight generator using normal uplink beamforming weights in accordance with the present invention;
FIG.16 is a graphic illustration of the downlink beamforming scheme using leaky uplink weights in accordance with embodiment 4 of the present invention;
FIG.17 illustrates one embodiment of downlink beamforming generator using leaky uplink weights in accordance

with the present invention;

DETAILED DESCRIPTION

[0025] FIG.1 shows the block diagram of downlink beamforming scheme according to prior art for improving downlink performance and capacity using base station antenna array. A plurality of mobile users share the same channels which can be a time slot for TDMA, a frequency band for FDMA or a set of spreading codes for CDMA. Using a plurality of transceivers, higher system capacity and better transmission performance can be achieved if proper uplink and downlink beamforming schemes are employed.

[0026] The prior art system first estimate each user's DOA values from the received uplink signals, then construct DCRs using downlink steering vectors for the estimated DOAs, finally set the DCRs as the downlink beamforming vectors. As discussed in the first section of this application, the prior art system is very complicated in the sense that all users' DOAs are to be estimated; also, this system cannot provide enough downlink capacity to match its uplink counterpart.

[0027] FIG.2 illustrates how the system and method of embodiment 1 of the present invention can overcome this problem. The received uplink signals are first used to estimate uplink channel covariance matrix (UCCM). UCCMs are then exploited to estimate downlink channel covariance matrix (DCCM), from which downlink beamforming weights are generated by inputting downlink data rate information.

[0028] Signals to be transmitted to mobile stations (MS) are finally weighted and combined for transmission through a plurality of transmitters. Therefore downlink beamforming weights can be generated from the received uplink signals directly, and no feedback or intermediate step for estimating DOAs is required.

[0029] UCCM can be estimated from received uplink signals directly, or from instantaneous uplink channel vector (IUCV) estimates derived using either pilot symbol assisted techniques or blind estimation techniques. For pilot symbol assisted approaches, both pilot symbols and decision-directed symbols can be used to improve estimation and tracking accuracy. For blind estimation techniques, constant modulus properties or finite alpha-beta properties of the modulated signals are exploited. Although there are some phase ambiguities within blind techniques, this will not affect DCCM estimates since they are blind to this ambiguity.

[0030] If multi-delay paths exist, each path's IUCVs can be estimated separately. This approach is applicable for wideband CDMA wireless communication systems.

[0031] Fig.3 shows the block diagram of DCCM estimator, which is one of the key parts of the present invention. According to Fig.3, DCCM is estimated by using UCCM through certain frequency calibration processing. Specifically, UCCM is first converted into a columnized UCCM vector, followed by an uplink-to-downlink transformer, followed by DCCM constructor. The theoretical basis of this method is given below. Here we consider a specific user.

- Uplink channel response is :

$$h_u(t) = \sum_{l=1}^{L} g(\theta_l)\alpha_{u,l}(t)a_u(\theta_l),$$

where g(θ) and $\alpha_{u,l}(t)$ are the antenna gain and uplink complex fading path strength respectively, $a_u(\theta_l)$ is the uplink steering vector at angle θ_l.

- For FDD system, according to reciprocal law, only the DOAs remain unchanged for uplink and downlink transmissions. Thus the downlink channel response for the same user is:

$$h_d(t) = \sum_{l=1}^{L} g(\theta_l)\alpha_{d,l}(t)a_d(\theta_l),$$

where $\alpha_{d,l}(t)$ is downlink complex fading path strength, $a_d(\theta_l)$ is the downlink steering vector at angle θ_l.

- UCCM:

$$R_u = \sum_{l=1}^{L} g^2(\theta_l)E[|\alpha_{u,}(t)|^2]a_u(\theta_l)a_u^H(\theta_l)$$

- DCCM:

$$R_d = \sum_{l=1}^{L} g^2(\theta_l) E[|\alpha_{d,l}(t)|^2] a_d(\theta_l) a_d^H(\theta_l)$$

- By considering $E[|\alpha_{u,l}(t)|^2] = E[|\alpha_{d,l}(t)|^2] = |\alpha_l|^2$, both uplink and downlink have the same spatial density function $\sigma(\theta) = g^2(\theta)|\alpha_l|^2$, which can be expressed in terms of Fourier series expansion over the possible DOA interval:

$$\sigma(\theta) = \sum_{k=-K}^{K} c(k) e^{-jkS\theta},$$

where S is the number of sectors per cell.

- $$R_u = \sum_{k=-K}^{K} c(k) Q_u^{(k)},$$

with

$$Q_u^{(k)} = \int_{-\pi/S}^{\pi/S} a_u(\theta) a_u^H(\theta) e^{-jkS\theta} d\theta,$$

or $Q_u c = r_u$, where $c = [c(-K), \Lambda\ c(k)]^T$,

$$Q_u = [q_u^{(-K)}, \Lambda\ q_u^{(k)}]$$

with $q_u^{(k)}$ and r_u being the columnized vectors of $Q_u^{(k)}$ and R_u.

- $$R_d = \sum_{k=-K}^{K} c(k) Q_d^{(k)},$$

with

$$Q_d^{(k)} = \int_{-\pi/S}^{\pi/S} a_d(\theta) a_d^H(\theta) e^{-jkS\theta} d\theta,$$

or $Q_d c = r_d$, where

$$Q_d = [q_d^{(-K)}, \Lambda\ q_d^{(k)}]$$

with $q_d^{(k)}$ and r_d being the columnized vectors of $Q_d^{(k)}$ and R_d.
- Thus linear relationship between elements of R_d and R_u is established: $r_d = A r_u$, where

$$A = Q_d (Q_u^H Q_u)^{-1} Q_u^H$$

is called frequency calibration (FC) matrix.
- The FC matrix A is only dependent on uplink and downlink carrier frequencies, transmit and receive array struc-

tures and cell sectorization, thus it can be computed and stored in advance and used directly during on-line processing

[0032] Physically, if the angular spread is small, the above technique maintains the same peak position for both uplink and downlink main beams generated from the principal eigenvectors of UCCM and DCCM, even though the system is a FDD system, thus is called peak constraint (PC) transform. This technique does not limit itself to the cases in which transmit antenna array structure is the same as receive antenna array structures. Here, the array structure means array geometry, antenna spacing and number of antenna elements.

[0033] Three algorithms for computing columnized DCCM vector can be used.

[0034] Algorithm 1 is applicable for any geometry of antenna array case, in which the FC matrix is a $n^2 \times m^2$ (possible complex) matrix, where m and n are the numbers of receive and transmit antenna elements.

$$\text{Algorithm 1 } r_d = Ar_u.$$

$r_d : n^2 \times 1,\ r_u : m^2 \times 1, A\colon n^2 \times m^2.$

[0035] Algorithm 2 is applicable for uniform linear array (ULA) in which the uplink and downlink channel covariance matrices are Hermitian and Toeplitz, thus only the first column and first row elements of UCCM R_u are used to construct those of DCCM R_d. Therefore, the associated FC matrix is a $(2n - 1) \times (2m - 1)$ real matrix.

$$\text{Algorithm 2 } p_d = Bp_u.$$

$p_d : (2n - 1) \times 1,\ p_u : (2m - 1) \times 1,\ B\colon (2n - 1) \times (2m - 1)\,.$

[0036] Algorithm 3 is also applicable for uniform linear array (ULA) in which the uplink and downlink channel covariance matrices are Hermitian and Toeplitz. Here, one n x m real FC matrix and one $(n - 1) \times (m - 1)$ real FC matrix are involved.

$$\text{Algorithm 3 } q_{d,r} = C_r q_{u,r},\ q_{d,i} = C_i q_{u,i}$$

$q_{d,r} : n \times 1,\ q_{u,r} : m \times 1,\ C_r : n \times m;$
$q_{d,i} : (n - 1) \times 1,\ q_{u,i} : (m - 1) \times 1,\ C_i : (n - 1) \times (m - 1)\,.$

[0037] As an example, for 6-elements ULA, 3 sectors/cell, and $f_u = 1.8GHz$ and $f_d = 2.0GHz$, the FC matrices are given below.

$$C_r = \begin{bmatrix} 1.0000 & 0 & 0 & 0 & 0 & 0 \\ -0.1031 & 1.0415 & 0.0778 & -0.0221 & 0.0081 & -0.0030 \\ 0.1136 & -0.2734 & 1.0100 & 0.1857 & -0.0474 & 0.0156 \\ -0.1371 & 0.2890 & -0.3613 & 0.9663 & 0.2890 & -0.0595 \\ 0.1950 & -0.3938 & 0.4112 & -0.4760 & 0.9377 & 0.3734 \\ -0.4983 & 0.9853 & -0.9544 & 0.9143 & -0.9056 & 1.0823 \end{bmatrix}$$

$$C_i = \begin{bmatrix} 0.9370 & 0.1406 & -0.0604 & 0.0299 & -0.0150 \\ -0.1223 & 0.9076 & 0.2526 & -0.0872 & 0.0384 \\ 0.0852 & -0.2139 & 0.8659 & 0.3507 & -0.0957 \\ -0.0851 & 0.1785 & -0.3126 & 0.8349 & 0.4351 \\ 0.1608 & -0.3126 & 0.4521 & -0.6091 & 0.9006 \end{bmatrix}$$

[0038] The computational complexities of the above three algorithms for estimating DCCM are different. Specifi-

cally, for $m = n = 6$, the complexity of Algorithm 3 is 25% of that of Algorithm 2, and 2.5% of that of Algorithm 1.

[0039] According to the present invention, TDD is a special case of FDD, in which uplink and downlink carrier frequencies are the same. However, the peak constraint transform can still take care of the possibly different receive and transmit antenna array structures.

[0040] Fig.4 shows first embodiment of downlink beamforming weight generator using DCCM. According to Fig.4, the principal eigenvector of DCCM can be used as the downlink beamforming weight vector. The functionality of the embodiment is the same as the prior art shown in Fig. 1, i.e., keeping the main beam of the downlink beam pattern toward to the desired user. However, the new embodiment is much simpler than the prior art shown in Fig. 1 as it is not involved with DOA estimation.

[0041] Fig.5 illustrates second embodiment of downlink beamforming weight generator using DCCM, in which downlink data rate information is used as well. According to the present invention, in order further to improve the performance of systems using base station antenna array, downlink data rate information can be taken into consideration in designing downlink beamforming weights. As an example, we will consider a DS-CDMA system with base station antenna array.

[0042] Suppose N mobile users share the same sector, and

$$h_{d,k}^{(l)}, l = 1, \mathrm{K}, L_k$$

are the lth path channel responses from the base station antenna array to the kth user. We consider per-user-per-weight downlink beamforming scheme due to its robustness to path changing problem and simplicity for implementation, and let $w_{d,k}$ denote the beamforming weight vector for user k. It can be shown that the instantaneous SIR at the Rake combiner output is given by $SIR_{d,k} = \tilde{S}_k / \tilde{I}_k$, where

$$\tilde{S}_k = \frac{P_{d,k}T}{N_0} \sum_{l=1}^{M} \left| w_{d,k}^H h_{d,k}^{(l)} \right|^2,$$

$$\tilde{I}_k = \frac{P_{d,k}T}{N_0} \frac{r_d(k)}{G} \left(\frac{\sum_{l=1}^{M} \sum_{i=1, i \neq l}^{L_k} \left| w_{d,k}^H h_{d,k}^{(i)} \right|^2 \left| w_{d,k}^H h_{d,k}^{(l)} \right|^2}{\sum_{l=1}^{M} \left| w_{d,k}^H h_{d,k}^{(l)} \right|^2} \right) + \sum_{j=1, j \neq k}^{N} \frac{r_d(j)}{G} \frac{P_{d,j}T}{N_0} \left(\sum_{l=1}^{L_k} \left| w_{d,j}^H h_{d,k}^{(l)} \right|^2 - \frac{\sum_{l=1}^{M} \left| w_{d,k}^H h_{d,k}^{(l)} \right|^2 \left| w_{d,j}^H h_{d,k}^{(l)} \right|^2}{\sum_{l=1}^{M} \left| w_{d,k}^H h_{d,k}^{(l)} \right|^2} \right) + 1]$$

with G being the processing gain, $P_{d,k}T$ the average signal energy-per-bit of one code channel user k, N_0 the one-sided spectrum density of AWGN noise, and $r_d(k)$ the normalized data rate of user k.

[0043] For DS-CDMA systems, power control is needed in order to compensate near-far problem. Specifically, for SIR-based power control, we try to maintain $SIR_{d,k} = \gamma_0$, where γ_0 is the target SIR threshold, or

$$(I - \gamma_0 F_d) p_d = g_d \qquad (1)$$

where

$$p_d = [P_{d,1}T, \mathrm{K}, P_{d,N}T]^T,$$

$$g_d = \left[\frac{\gamma_0 N_0}{\sum_{l=1}^{L_1} \left| w_{d,1}^H h_{d,1}^{(l)} \right|^2}, \mathrm{K}, \frac{\gamma_0 N_0}{\sum_{l=1}^{L_N} \left| w_{d,N}^H h_{d,N}^{(l)} \right|^2} \right]^T,$$

and $F_d = DFR$ with

$$D = diag\left[\frac{1}{\sum_{l=1}^{L_1} \left| w_{d,1}^H h_{d,1}^{(l)} \right|^2}, \mathrm{K}, \frac{1}{\sum_{l=1}^{L_N} \left| w_{d,N}^H h_{d,N}^{(l)} \right|^2} \right],$$

$$R = diag[r_d(1),...,r_d(N)],$$

$$[F]_{i,j} = \begin{cases} \dfrac{1}{G}\left(\sum_{l=1}^{L_i} \left| w_{d,i}^H h_{d,i}^{(l)} \right|^2 - \dfrac{\sum_{l=1}^{L_i} \left| w_{d,i}^H h_{d,i}^{(l)} \right|^4}{\sum_{l=1}^{M} \left| w_{d,i}^H h_{d,i}^{(l)} \right|^2} \right), & i = j \\ \dfrac{1}{G}\left(\sum_{l=1}^{L_i} \left| w_{d,j}^H h_{d,i}^{(l)} \right|^2 - \dfrac{\sum_{l=1}^{L_i} \left| w_{d,i}^H h_{d,i}^{(l)} \right|^2 \left| w_{d,j}^H h_{d,i}^{(l)} \right|^2}{\sum_{l=1}^{M} \left| w_{d,j}^H h_{d,i}^{(l)} \right|^2} \right), & i \neq j \end{cases} \quad (2)$$

[0044] Given downlink beamforming weights, if we do not consider power constraint, then

$$\frac{1}{\rho(F_d)}$$

is actually the maximum achievable SIR threshold, where $\rho(F_d)$ is the spectral radius of F_d. The outage probability is defined as

$$P_{out} = \Pr\left(\frac{1}{\rho(F_d)} < \gamma_{d,0}\right),$$

where $\gamma_{d,0}$ is the target SIR threshold. Therefore, the objective of downlink beamforming is, for given $\gamma_{d,0}$, to choose a set of beamforming weights $w_{d,k}$'s, such that the outage probability is minimal, or maximum number of users can be supported within the same sector. As the outage probability is most probably affected by the cases whose

$$\frac{1}{\rho(F_d)}$$

value is near $\gamma_{d,0}$. The objective is equivalent to finding a set of weights such that minimum total transmitted power is required in order for all users to achieve the SIR requirement, $\gamma_{d,0}$. This is, obviously, a difficult multi-variable optimization problem. Fortunately, this problem can be converted into an easily solved problem by making some approximations. Specifically, since the optimal weight vector, $w_{d,i}$, generates almost equal beam responses at the DOAs of all paths of user i, we have

$$[\bar{F}]_{i,j} \approx \frac{L_i\text{-}1}{GL_i}\left(\sum_{l=1}^{L_i} \left|w_{d,j}^H h_{d,i}^{(l)}\right|^2\right), \text{ for } i \neq j, \tag{3}$$

[0045] In this case, specifically, define $F_u = D\bar{F}^T R$, then $\rho(F_d) \approx \rho(F_u)$ since both D and R are diagonal matrices. Note F_u can be considered as a virtual uplink matrix, which is an uplink counterpart of F_d. Thus the solution to the above problem can be obtained using iterative virtual power weighted (IVPW) algorithm as shown in Fig.5, or more simply, virtual power weighted (VPW) algorithm illustrated in Fig.6, or spatial distribution weighted (SDW) algorithm shown in Fig.7. Here, for simplicity, we use an equivalent-one-path channel vector (EOCV), $h_{d,k}$, to replace multi-delay paths of user k.

[0046] Fig.5 shows IVPW algorithm for estimating downlink beamforming weights according to the present invention. It consists of the following iterative steps.

(1.1) Input downlink data rate information (DDRI), $r_d(k)$, for $k = 1,\Lambda\ , N$.
(1.2) Choose initial weight vectors, say $w_{d,k} = h_{d,k}$, for $k = l,\ldots,N$;
(1.3) Determine virtual uplink power vector for given weight vectors: $p_v = \gamma_{d,0}(I - \gamma_{d,0}F)^{-1}g$;
(1.4) For given power vector, derive weight vectors using maximum SINR criteria, for all users, for $k = l,\ldots,N$,

$$\max \frac{P_{v,k} w_{d,k}^H R_{d,k} w_{d,k}}{w_{d,k}^H \left(\sum_{j \neq k}^{N} \frac{1}{G} R_{d,j} r_d(j) P_{v,j} + r_d(k) P_{v,k} R_{d,k} + N_0 I\right) w_{d,k}};$$

(1.5) Update (1.3) and (1.4) until power and weight vectors are converged. The converged weight vectors are used as downlink beamforming weight vectors.

[0047] IVPW algorithm involves iterative updates. Fig.6 shows third embodiment for generating downlink beamforming weights according to the present invention. This embodiment is called VPW algorithm, in which no iterative update process is required. According to Fig.6, the embodiment consists of the following steps.

(2.1) Input downlink data rate information (DDRI), $r_d(k)$, for $k = 1,\Lambda\ , N$.
(2.2) Choose initial weight vectors: $w_{d,k} = h_{d,k}$, for $k = l,\ldots,N$;
(2.3) Determine virtual uplink powervector: $p_v = \gamma_{d,0}(I - \gamma_{d,0}F_0)^{-1}g_0$;
(2.4) Derive weight vectors using maximum SINR criteria, for all users,

$$\max \frac{P_{v,k} w_{d,k}^H R_{d,k} w_{d,k}}{w_{d,k}^H \left(\sum_{j \neq k}^{N} \frac{1}{G} R_{d,j} r_d(j) P_{v,j} + r_d(k) P_{v,k} R_{d,k} + N_0 I \right) w_{d,k}};$$

(2.4) The above solution is set for downlink beamforming weight vector.

[0048] Further simplification is derived in the fourth embodiment of downlink beamforming weight generator, SDW algorithm, according to the present invention, which is shown in Fig.7. According to Fig.7, we may simplify the power vector computation by replacing matrix inverse with an approximation. Specifically, the new embodiment consists of the following step.

(3.1) Input downlink data rate information (DDRI), $r_d(k)$, for $k = 1, \Lambda, N$.
(3.2) Choose initial weight vectors: $w_{d,k} = h_{d,k}$, for $k = I, ..., N$;
(3.3) Determine virtual uplink powervector: $p_v = \gamma_{d,0}(I + \gamma_{d,0} F_0) g_0$;
(3.4) Derive weight vectors using maximum SINR criteria, for all users,

$$\max \frac{P_{v,k} w_{d,k}^H R_{d,k} w_{d,k}}{w_{d,k}^H \left(\sum_{j \neq k}^{N} \frac{1}{G} R_{d,j} r_d(j) P_{v,j} + r_d(k) P_{v,k} R_{d,k} + N_0 I \right) w_{d,k}};$$

(3.5) The above solution is set for downlink beamforming weight vector.

[0049] In Fig.5 - Fig.7, although we don't have downlink channel information in determining virtual power vector, we replace it with stable downlink channel vector (SDCV) estimate, which is defined as the principal eigenvector of DCCM.

[0050] Fig.8 illustrates the block diagram of downlink beamforming scheme using uplink channel estimates in accordance with embodiment 2 of the present invention. The received uplink signals are first used to estimate instantaneous uplink channel vectors (IUCVs), which are then passed to downlink channel estimator, followed by beamforming weight generator. Downlink data rate information is also added in generating downlink beamforming weights. Signals to be transmitted to mobile users are finally weighted by these weights and combined for transmission through a plurality of transmitters.

[0051] Fig.9 shows one embodiment of downlink channel estimator according to the present invention. IUCVs are first used to calculate UCCM via time average approach. DCCM is estimated using peak constraint method. The principal eigenvector is used as SDCV estimate.

[0052] Fig.10 illustrates another embodiment of downlink channel estimator according to the present invention. Different from peak constraint method, the main idea of this estimator is to keep same the null positions of beams generated from SUCVs and SDCVs. Thus we call this as null constraint method.

[0053] Specifically, null constraint method is described as follows.

(4.1) Use IUCVs to calculate UCCM;
(4.2) Choose principal eigenvector of UCCM as SUCV, $h_{u,k}$
(4.3) Determine uplink beam nulls $z_{u,k}^{(i)}$ from the polynomial formed from SUCV:

$$h_{u,k}^{(1)}(1 - z_{u,k}^{(1)} z^{-1}) \Lambda (1 - z_{u,k}^{(M-1)} z^{-1}) = \sum_{i=1}^{M} h_{u,k}^{(i)} z^{-i+1};$$

(4.4) Transform the phase components of all the uplink beam pattern nulls $z_{u,l}$ into their downlink counterpart:

$$\varphi_{d,k}^{(i)} = \varphi_{u,k}^{(i)} f_d / f_u$$

where

$$z_{u,k}^{(i)} = A_i e^{j\phi_{u,k}^{(i)}};$$

(4.5) Construct the downlink beam nulls $z_{d,k}^{(i)}$:

$$z_{d,k}^{(i)} = A_i e^{j\phi_{d,k}^{(i)}};$$

(4.6) Construct downlink polynomial:

$$h_{d,k}^{(1)}(1 - z_{d,k}^{(1)} z^{-1}) \Lambda (1 - z_{d,k}^{(M-1)} z^{-1}) = \sum_{i=1}^{M} h_{d,k}^{(i)} z^{-i+1},$$

and determine SDCVs by choosing the coefficients of the constructed polynomial.

[0054] Similar to IVPW, VPW and SPW algorithms with DCCMs as inputs, shown in Figs.5-7, Figs.11-13 illustrate IVPW, VPW and SPW algorithms for generating downlink beamforming weights using SDCVs as input, respectively.

[0055] Fig.14 shows downlink beamforming scheme using normal uplink beamforming weights in accordance with embodiment 3 of the present invention. According to this embodiment, downlink beamforming weights can be generated by direct modifying uplink weights via null constraint method. As uplink beamforming weights are already at hand, the beauty of this embodiment is its simplicity in terms of software and hardware complexities.

[0056] Fig.15 shows downlink beamforming weight generator based on null constraint method. In fact, uplink beamforming weights are optimal for uplink reception. If the system is time division duplex (TDD), then uplink weights can be used for downlink directly, since uplink beam pattern is the same as downlink beam pattern right here. However, for FDD systems, if uplink weights are used for downlink transmission directly, the null positions and main beam position will be shifted due to different receive and transmit carrier frequencies. Null constraint method designs downlink beamforming weights such that the null positions of downlink beam patters are kept the same as those of the uplink one.

[0057] Specifically, null constraint method is described as follows.

(5.1) Determine the uplink beam pattern nulls $z_{u,k}^{(i)}$ from the polynomial formed from uplink weight:

$$w_{u,k}^{(1)}(1 - z_{u,k}^{(1)} z^{-1}) \Lambda (1 - z_{u,k}^{(M-1)} z^{-1}) = \sum_{i=1}^{M} w_{u,k}^{(i)} z^{-i+1};$$

(5.2) Transform the phase components of all the uplink beam pattern nulls $z_{u,i}$:

$$\varphi_{d,k}^{(i)} = \varphi_{u,k}^{(i)} f_d / f_u$$

where

$$z_{u,k}^{(i)} = A_i e^{j\phi_{u,k}^{(i)}};$$

(5.3) Construct the downlink beam pattern nulls $z_{d,k}^{(i)}$:

$$z_{d,k}^{(i)} = A_i e^{j\phi_{d,k}^{(i)}};$$

(5.4) Construct the downlink polynomial:

$$w_{d,k}^{(1)}(1 - z_{d,k}^{(1)}z^{-1}) \Lambda (1 - z_{d,k}^{(M-1)}z^{-1}) = \sum_{i=1}^{M} w_{d,k}^{(i)} z^{-i+1},$$

and choose the coefficients of the constructed polynomial as downlink beamforming weights.

[0058] Fig.16 shows the downlink beamforming scheme using leaky uplink beamforming weights in accordance with embodiment 4 of the present invention.

[0059] Fig.17 shows downlink beamforming weight generator using uplink leaky MMSE (LMMSE) weights. This approach generates downlink beamforming weights by modifying uplink LMMSE weights together with peak constraint method.

[0060] Usually, leaky MMSE is used to provide robust uplink beamforming weights. Here, another property of leaky MMSE - beam adjustment, is exploited. As an embodiment, we use a CDMA system to describe the effect of leaky factor. Suppose uplink uses per-user-per-weight beamforming scheme. The cost function for estimating uplink weights is given by

$$J = |w^H Y\eta - d|^2 + \alpha \| Y\eta \|^2 \| w \|^2,$$

where w is the beamforming weight vector (common for all delay paths), $Y = [y_1, K, y_L]$ with y_i being the despread signal vector across the array element in the ith delay path, $\eta = [\eta_1, K, \eta_L]^T$ is the RAKE coefficient vector, d is the training symbol, and α is leaky factor.

[0061] The normalized leaky LMS update equation is given by

$$w(k+1) = (1 - \mu\alpha)w(k) - \mu \frac{z(k)e^*(k)}{z^H(k)z(k)},$$

where $z(k) = Y(k)\eta(k)$, is the composite beamformer input, $e(k) = w^H(k)z(k) - d(k)$, is the error signal. Similarly, we can obtain leaky RLS update equation.

[0062] LMMSE is a generalization of the normal MMSE (NMMSE) algorithm, which corresponds to $\alpha = 0$. When α is large enough (smaller than the maximum allowed value, same as below), the generating weights are composite uplink channel responses, or so-called MRC weights. LMMSE provides a flexible leaky factor for adjusting uplink beam pattern. Specifically, when α is zero, the generated beam pattern simultaneously takes care of the desired user's antenna responses as well as suppression to interference. When α becomes larger, however, the main beam will be getting closer and closer to the desired user's direction, while less consideration will be paid to the suppression to interference. Obviously, for uplink reception, NMMSE gives the best performance and maximum system capacity; and the larger α is, the worse uplink performance becomes.

[0063] Here, we are interested in modifying uplink weights for downlink use. According to the present invention, optimal downlink beamforming weights can be generated from uplink LMMSE weights with a moderate leaky factor together with some frequency calibration processing, such as peak constraint algorithm.

[0064] If uplink NMMSE weights are used for downlink, with or without peak constraint transform, although the interference can be suppressed in some extent, the desired user's antenna responses will not be well taken care of. Specifically, in some extreme cases, the desired users may fall into the null positions or side lobes of their own beam patterns. This is because that the difference between uplink and downlink carrier frequencies can be as high as 10% or even 20% of the uplink carrier frequency, and that the main beam of the beam pattern generated from the NMMSE weights are usually biased from the actual nominal DOAs, especially when two or more wireless users are spatially closed.

[0065] When a LMMSE with large leaky factor is used, a MRC weight vector is generated, whose beam pattern's main beam will direct toward the desired user, if peak constraint transform is added. However, no considerations are paid for interference suppression in this case.

[0066] Using peak constraint method, downlink weights obtained by modifying uplink LMMSE weights with moderate leaky factor simultaneously suppress interference in some extent, and form the main beam near the desired user's direction. Keeping in mind that optimal uplink beamforming weights take care of both desired user's antenna responses as well as suppression to interference, we may conclude that LMMSE with moderate leaky factor together with peak constraint method gives optimal downlink beamforming weights.

[0067] While the above description contains certain specifications, these should not be considered as limitations on the scope of the invention, but rather as an exemplification of one preferred embodiment and application thereof. It will be apparent to those skilled in the art that various modifications can be made to the downlink beamforming scheme revealed in the present invention without departing from the scope and sprit of the invention. It is intended that the

present invention cover modifications and variations of the systems and methods which are from the scope of the appended claims and equivalents.

Claims

1. A method for downlink capacity enhancement in a wireless communications system comprising a base station with antenna array and terminals that are physically remote from said base station, the method comprising steps of:

   receiving at said base station antenna array combinations of arriving signals from said plurality of remote terminals;
   estimating an uplink channel covariance matrix (UCCM) for each of said terminals from said combinations of arriving signals;
   constructing from each of said UCCM a downlink channel covariance matrix (DCCM);
   inputting downlink data rate information (DDRI);
   calculating from all said DCCM and DDRI a downlink weight vector for each of said terminals; and
   transmitting a set of information signals from said base station antenna array according to said downlink weight vectors.

2. The method of Claim 1 wherein the estimating step comprises:

   forming from said combinations of arriving signals an uplink channel vector for each of said terminals;
   establishing a UCCM for each said remote terminal by taking a linear combination of outer products of the corresponding the uplink channel vectors.

3. The method of Claim 2 wherein the forming step comprises:

   calculating from said combinations of arriving signals and sets of uplink training sequences associated with said remote terminals an uplink minimum mean-square-error (MMSE) weight vector for each of said terminals;
   assigning said uplink MMSE weight vector as the uplink channel vector for each of said terminals.

4. The method of Claim 1 wherein the plurality of remote terminals are CDMA terminals, each of which has an unique PN code sequence.

5. The method of Claim 4 wherein the estimating step comprises:

   forming a despread signal for each of said terminals from said combinations of arriving signals and said associated PN code sequence; and
   establishing a UCCM for each said remote terminal by taking a linear combination of outer products of the corresponding despread signal.

6. The method of Claim 5 wherein the establishing step comprises:

   computing an uplink channel vector for each of said terminals from the associated despread signal and at least one training sequence associated with each remote terminal; and
   constructing a UCCM for each said remote terminal by taking a linear combination of outer products of the corresponding uplink channel vector,

7. The method of Claim 6 wherein the computing step comprises:

   calculating an estimated gradient of the error function that includes weighted magnitude square of said uplink channel vector

   $$|h^H y - d|^2 + \alpha|h|^2$$

   where h is the uplink channel vector, y the despread signal, d the training sequence, α a weighting constant; and
   updating said uplink channel vector by adjusting it according to said estimated gradient.

8. The method of Claim 5 wherein the establishing step comprises:

computing an uplink weight vector for each of said terminals from the associated despread signal; and
constructing a UCCM for each said remote terminal by taking a linear combination of outer products of the corresponding uplink weight vector.

9. The method of Claim 8 wherein the computing step comprises:

calculating an estimated gradient of the error function that includes weighted magnitude square of said MMSE weight vector

$$|w^H y - d|^2 + \alpha |w|^2$$

where w is the uplink MMSE weight vector, y the despread signal, d the training sequence, α a weighting constant; and
updating said uplink MMSE weight vector by adjusting it according to said estimated gradient.

10. The method of Claim 1 wherein the constructing step comprises the substeps of:

columnising the said UCCM to a first column vector;
calculating a second column vector by multiplying a frequency calibration matrix M_A (FCM-M_A) with said first column vector, the FCM-$M_{A,}$, a $n^2 \, x \, m^2$ matrix where m and n are the number of receive and transmit antenna elements, being only dependent on the carrier frequencies, transmit and receive array structures and cell sectorisation; and
constructing said DCCM from said second column vector.

11. The method of Claim 1 wherein the UCCM is used as the DCCM in the constructing step.

12. The method of Claim 1 wherein the constructing step comprises the substeps of:

extracting from the first column and first row of said UCCM to form a first column vector;
calculating a second column vector by multiplying a frequency calibration matrix M_B (FCM-M_B) with said first column vector, the FCM-M_B, a $(2n\text{-}l) \, x \, (2\text{m}\text{-}l)$ matrix where m and n are the number of receive and transmit antenna elements, being only dependent on the carrier frequencies, transmit and receive array structures and cell sectorisation; and
constructing said DCCM from said second column vector.

13. The method of Claim 1 wherein the constructing step comprises the substeps of:

extracting from the first column and first row of said UCCM to form a first column vector;
extracting the real part of said first column vector to form a second column vector and the imaginary part of said first column vector to form a third column vector;
calculating a fourth column vector by multiplying a frequency calibration matrix M_c (FCM-M_c) with said second column vector, the FCM-M_c , a $n \, x \, m$ matrix where m and n are the number of receive and transmit antenna elements, being only dependent on the carrier frequencies, transmit and receive array structures and cell sectorisation;
calculating a fifth column vector by multiplying a frequency calibration matrix M_D (FCM-M_D) with said third column vector, the FCM-M_D , a $(n\text{-}l) \, x \, (m\text{-}l)$ matrix where m and n are the number of receive and transmit antenna elements, being only dependent on the carrier frequencies, transmit and receive array structures and cell sectorisation;
forming a complex sixth column vector with real part being said fourth column vector and imaginary part being said fifth column vector; and
constructing said DCCM from said sixth column vector.

14. The method of Claim 1 wherein the downlink weight vector for each of said terminals is the dominant eigenvector of the said DCCM corresponding to the said terminals.

15. The method of Claim 1 wherein the calculating step comprises the substeps of:

calculating a channel vector for each of said terminals by taking the dominant eigenvector of the corresponding DCCM; and

repeating the steps of:

determining a set of power coefficients from a set of downlink system parameters concerning all mobile terminals that include said downlink weight vectors, said channel vector, downlink information data transmission rate and downlink link quality requirement of each of said terminals;

computing an autocorrelation matrix by taking a weighted sum of all DCCM corresponding to said terminals according to said set of power coefficients; and forming a downlink weight vector for each of said terminals from said autocorrelation matrix and corresponding DCCM for said terminal, wherein the downlink weight vector has maximal projection onto the corresponding DCCM and minimal projection onto said autocorrelation matrix

until the said set of power coefficients and downlink weight vectors have converged.

16. The method of Claim 1 wherein the calculating step comprises the substeps of:

calculating a channel vector for each of said terminals by taking the dominant eigenvector of the corresponding DCCM;

determining a set of power coefficients from a set of downlink system parameters concerning all mobile terminals that include said channel vector, downlink information data transmission late and downlink link quality requirement of each of said terminals;

computing an autocorrelation matrix by taking a weighted sum of all DCCM corresponding to said terminals according to said set of power coefficients; and

forming a downlink weight vector for each of said terminals from said autocorrelation matrix and corresponding DCCM for said terminal, wherein the downlink weight vector has maximal projection onto the corresponding DCCM and minimal projection onto said autocorrelation matrix.

17. The method of Claim 15 or 16 wherein said downlink weight vector for each of said terminals is the dominant generalised eigenvector of the corresponding DCCM and said autocorrelation matrix.

18. The method of Claim 15 or 16 wherein said downlink weight vector for each of said terminals is the dominant eigenvector of a matrix, which is the product of the inverse of said autocorrelation matrix and the corresponding DCCM.

19. The method of Claim 15 and 16 wherein said downlink weight vector for each of said terminals is the product of the inverse of said autocorrelation matrix and the said corresponding channel vector.

20. A method for downlink capacity enhancement in a wireless communications system comprising a base station with antenna array and terminals that are physically remote from said base station, the method comprising steps of:

receiving at said base station antenna array combinations of arriving signals from said plurality of remote terminals;

estimating an uplink weight vector for each of said terminals from said combinations of arriving signals;

constructing from each of said uplink weight vector a downlink weight vector; and

transmitting the set of information signals from said base station antenna array according to said downlink weight vectors.

21. The method of Claim 20 wherein the said uplink weight vector for each of said terminals is the corresponding uplink channel vector estimated from said combinations of arriving signals.

22. The method of Claim 20 wherein the plurality of remote terminals are CDMA terminals, each of which has an unique PN code sequence.

23. The method of Claim 22 wherein the estimating step comprises:

forming a despread signal for each of said terminals from said combinations of arriving signals and said associated PN code sequence; and

computing said uplink weight vector from corresponding despread signal.

24. The method of Claim 23 wherein the computing step comprises:

calculating a estimated gradient of the error function that includes weighted magnitude square of said uplink weight vector

$$|w^H y - d|^2 + \alpha|w|^2$$

where w is the uplink MMSE weight vector, y the despread signal, d the training sequence, α a weighting constant; and
updating said uplink MMSE weight vector by adjusting it according to said estimated gradient.

25. The method of Claim 20 wherein the constructing step comprises:

determining the zeros of the polynomial whose coefficients are the elements of the uplink weight vector;
forming new polynomial zeros by scaling the phase of said zeros by a factor that is related to the ratio of the downlink frequency to the uplink frequency; and
establishing said downlink weight vector by constructing a new polynomial using said new polynomial zeros and using the coefficients of said new polynomial as the elements of said downlink weight vector.

26. The method of Claim 1 wherein the uplink weight vector is used as the downlink weight vector in the constructing step.

27. A method for downlink capacity enhancement in a wireless communications system, comprising a base station with antenna array and terminals that are physically remote from said base station, the method comprising steps of:

receiving at said base station antenna array combinations of arriving signals from said plurality of remote terminals;
estimating an uplink channel vector for each of said terminals from said combinations of arriving signals;
constructing from each of said uplink channel vector a downlink channel vector;
calculating from all said downlink channel vector a downlink weight vector for each of said terminals; and
transmitting the set of information signals from said base station antenna array according to said downlink weight vectors.

28. The method of Claim 27 wherein the plurality of remote terminals are CDMA terminals, each of which has an unique PN code sequence.

29. The method of Claim 28 wherein the estimating step comprises:

forming a despread signal for each of said terminals from said combinations of arriving signals and said associated PN code sequence; and
computing said uplink channel vector from corresponding despread signal.

30. The method of Claim 29 wherein the computing step comprises:

calculating an estimated gradient of the error function that includes weighted magnitude square of said uplink channel vector

$$|h^H y - d|^2 + \alpha|h|^2$$

where h is the uplink channel vector, y the despread signal, d the training sequence, α a weighting constant; and
updating said uplink channel vector by adjusting it according to said estimated gradient.

31. The method of Claim 29 wherein the computing step comprises:

establishing a UCCM for each said remote terminal by taking a linear combination of outer products of the corresponding despread signal; and
forming the uplink channel vector for each said remote terminal by taking the dominant eigenvector of corresponding said UCCM.

32. The method of Claim 27 wherein the constructing step comprises:

determining the zeros of the polynomial whose coefficients are the elements of the uplink channel vector;
forming new polynomial zeros by scaling the phase of said zeros by a factor that is related to the ratio of the downlink frequency to the uplink frequency; and
establishing said downlink channel vector by constructing a new polynomial using said new polynomial zeros and using the coefficients of said new polynomial as the elements of said downlink channel vector.

33. The method of Claim 27 wherein the uplink channel vector is used as the downlink channel vector in the constructing step.

34. The method of Claim 27 wherein the downlink channel vector for each of said terminals is used as the corresponding downlink weight vector.

35. The method of Claim 27 wherein the calculating step comprises repeating the substeps of:

determining a set of power coefficients from a set of downlink system parameters concerning all mobile terminals that include said downlink weight vectors, said downlink channel vector, downlink information data transmission rate and downlink link quality requirement of each of said terminals;
computing an autocorrelation matrix by taking a weighted sum of the outer products of said downlink channel vectors corresponding to said terminals according to said set of power coefficients and downlink information data transmission rates; and
forming a downlink weight vector for each of said terminals by taking the product of the inverse of said autocorrelation matrix and corresponding said downlink channel vector
until the said set of power coefficients and down link weight vectors have converged.

36. The method of Claim 27 wherein the calculating step comprises the substeps of:

determining a set of power coefficients from a set of downlink system parameters concerning all mobile terminals that include said downlink weight vectors, said downlink channel vector, downlink information data transmission rate and downlink link quality requirement of each of said terminals;
computing an autocorrelation matrix by taking a weighted sum of the outer products of said downlink channel vectors corresponding to said terminals according to said set of power coefficients and downlink information data transmission rates; and
forming a downlink weight vector for each of said terminals by taking the product of the inverse of said autocorrelation matrix and corresponding said downlink channel vector.

37. A base station for a wireless communications system, the base station comprising:

an uplink receive antenna array for receiving arriving signals from a plurality of remote terminals on respective uplink channels;
an uplink weight generator for estimating a property of an uplink channel;
a downlink weight generator operable to derive downlink weights from the uplink channel property; and
a downlink transmit antenna array to transmit signals to the remote terminals in accordance with the desired downlink weights.

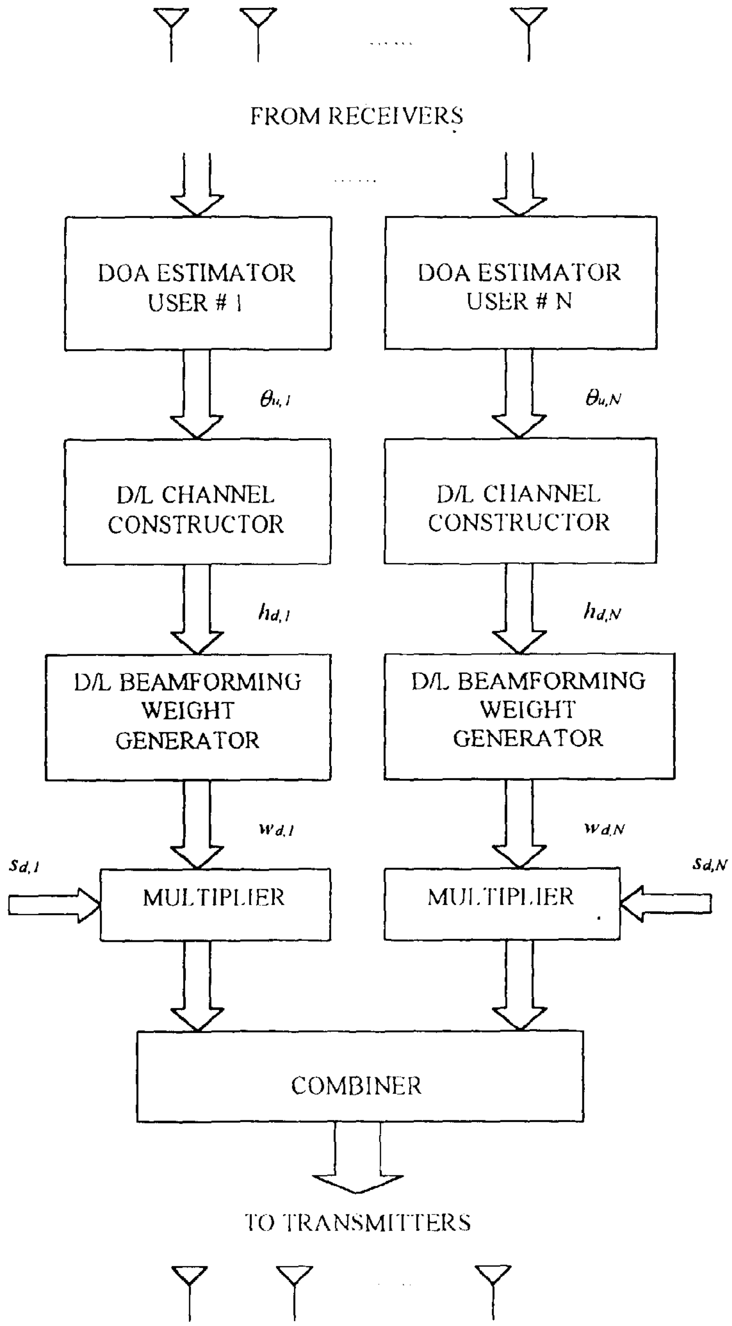
38. A base station according to Claim 37, wherein the uplink receive antenna array is the same as the downlink transmit antenna array.

39. A base station according to Claim 37, wherein the uplink receive antenna array is separate from the downlink transmit antenna array.

40. A base station according to any one of Claims 37 to 39, wherein the property of the uplink channel comprises at least one of the uplink channel covariance matrices, the uplink channel responses or the uplink beamforming weights.

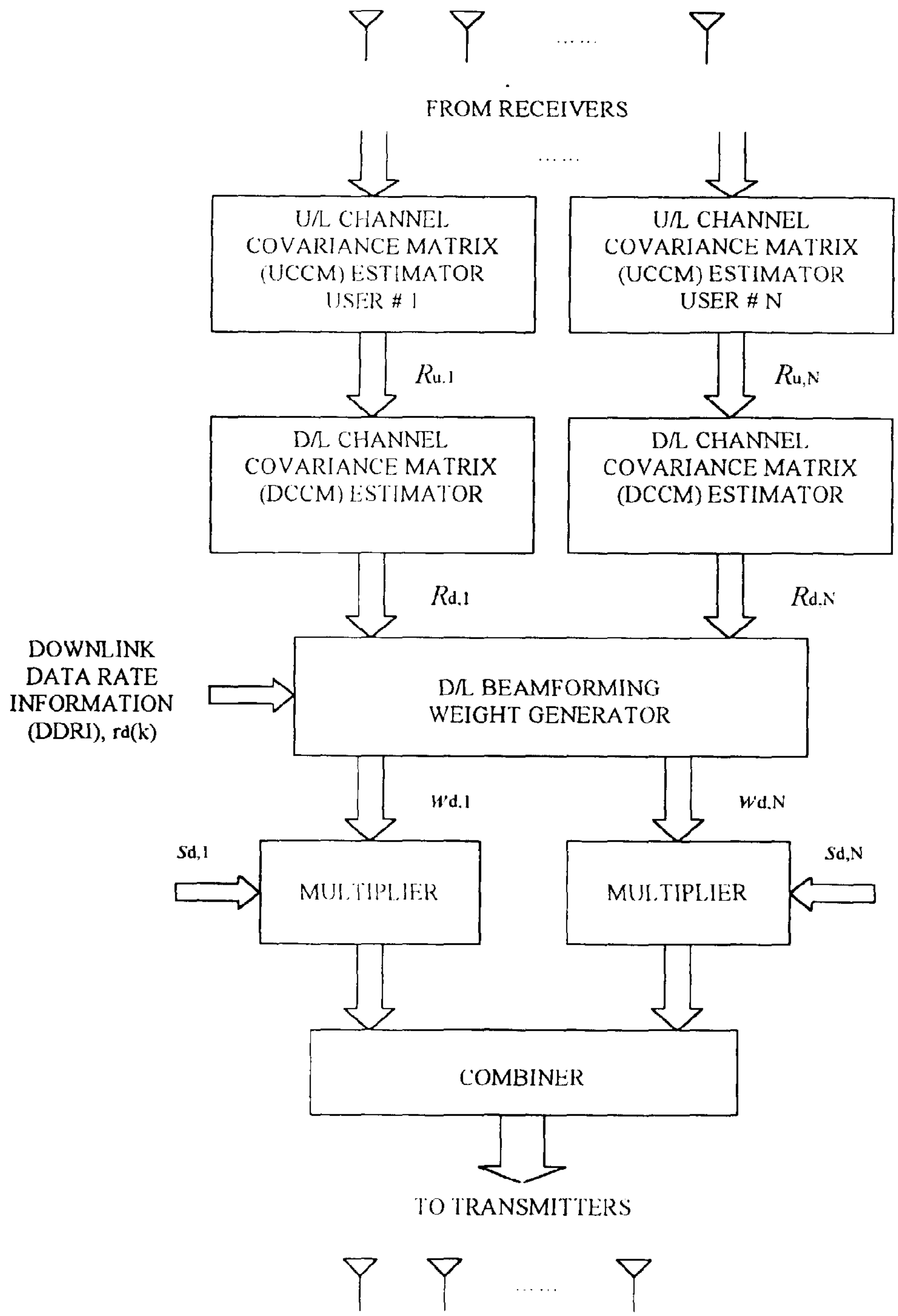
41. A base station according to any one of Claims 37 to 40, wherein uplink spatial de-multiplexing means and downlink spatial multiplexing means are provided.

42. A communication system incorporating a base station according to any one of Claims 37 to 41 and a plurality of remote terminals.



PRIOR ART
FIG. 1

FIG. 2


FROM RECEIVERS
U/L CHANNEL COVARIANCE MATRIX (UCCM) ESTIMATOR USER # 1
U/L CHANNEL COVARIANCE MATRIX (UCCM) ESTIMATOR USER # N
$R_{u,1}$
$R_{u,N}$
D/L CHANNEL COVARIANCE MATRIX (DCCM) ESTIMATOR
D/L CHANNEL COVARIANCE MATRIX (DCCM) ESTIMATOR
$R_{d,1}$
$R_{d,N}$
DOWNLINK DATA RATE INFORMATION (DDRI), rd(k)
D/L BEAMFORMING WEIGHT GENERATOR
$W_{d,1}$
$W_{d,N}$
$S_{d,1}$
$S_{d,N}$
MULTIPLIER
MULTIPLIER
COMBINER
TO TRANSMITTERS

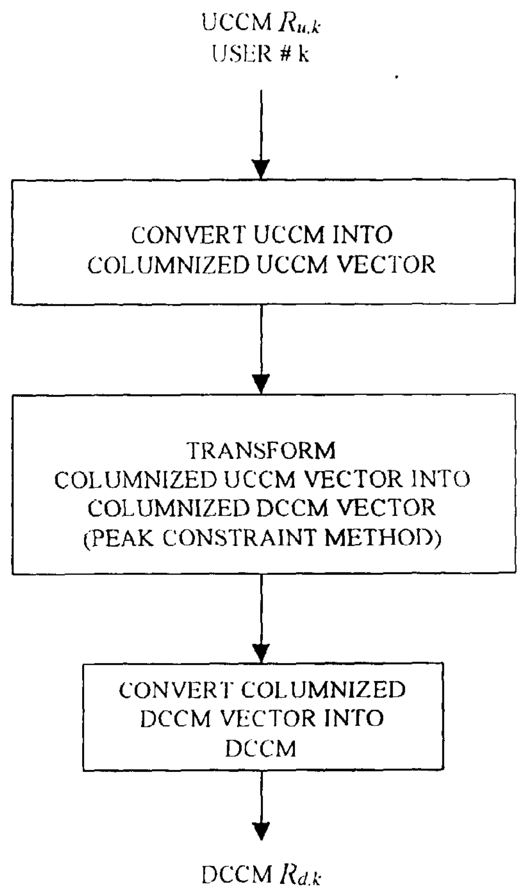
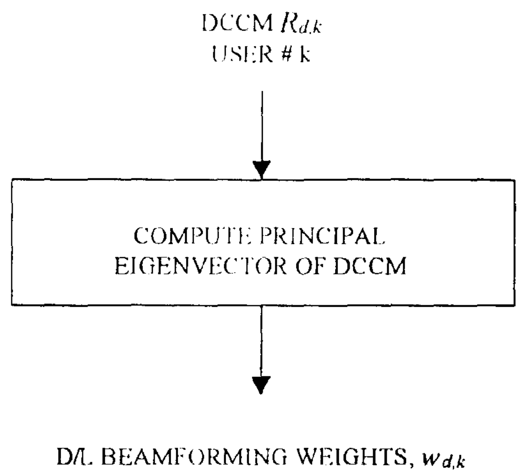

UCCM $R_{u,k}$
USER # k
CONVERT UCCM INTO COLUMNIZED UCCM VECTOR
TRANSFORM COLUMNIZED UCCM VECTOR INTO COLUMNIZED DCCM VECTOR (PEAK CONSTRAINT METHOD)
CONVERT COLUMNIZED DCCM VECTOR INTO DCCM
DCCM $R_{d,k}$

FIG. 3

FIG. 4


DCCM $R_{d,k}$
USER # k
COMPUTE PRINCIPAL EIGENVECTOR OF DCCM
D/L BEAMFORMING WEIGHTS, $w_{d,k}$

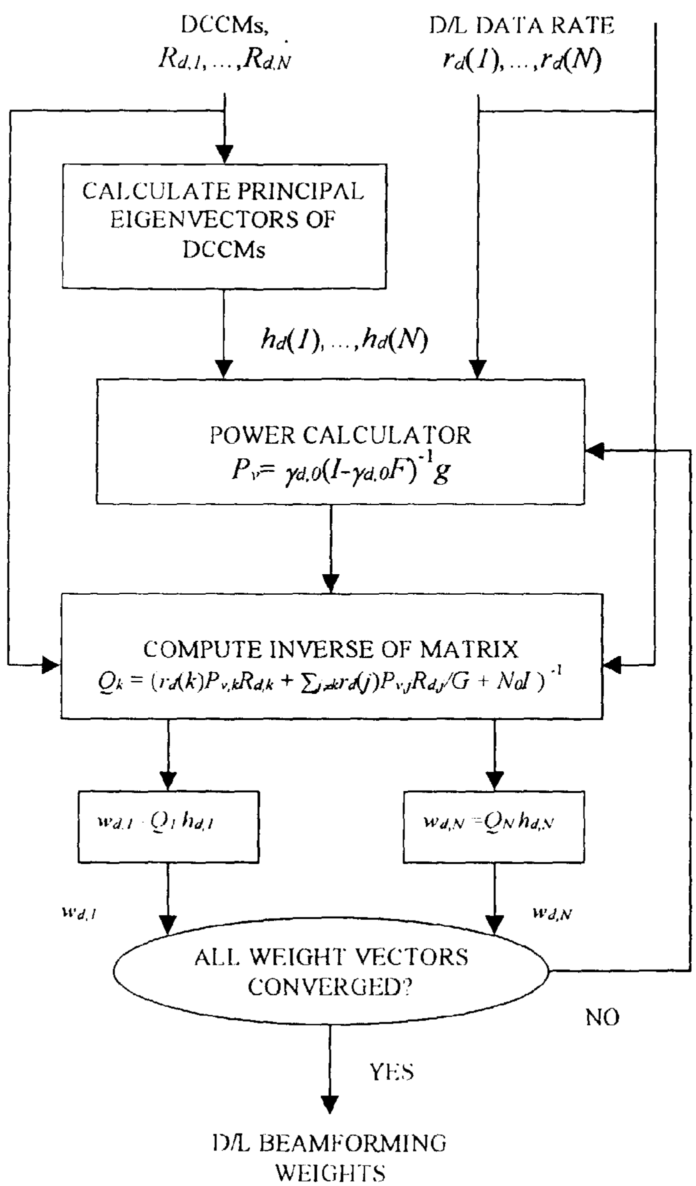

DCCMs,
$R_{d,1},\ldots,R_{d,N}$
D/L DATA RATE
$r_d(1),\ldots,r_d(N)$
CALCULATE PRINCIPAL EIGENVECTORS OF DCCMs
$h_d(1),\ldots,h_d(N)$
POWER CALCULATOR
$P_v = \gamma_{d,0}(I-\gamma_{d,0}F)^{-1}g$
COMPUTE INVERSE OF MATRIX
$Q_k = (r_d(k)P_{v,k}R_{d,k} + \sum_{j\neq k} r_d(j)P_{v,j}R_{d,j}/G + N_0 I)^{-1}$
$w_{d,1} = Q_1 h_{d,1}$
$w_{d,N} = Q_N h_{d,N}$
$w_{d,1}$
$w_{d,N}$
ALL WEIGHT VECTORS CONVERGED?
NO
YES
D/L BEAMFORMING WEIGHTS

FIG. 5

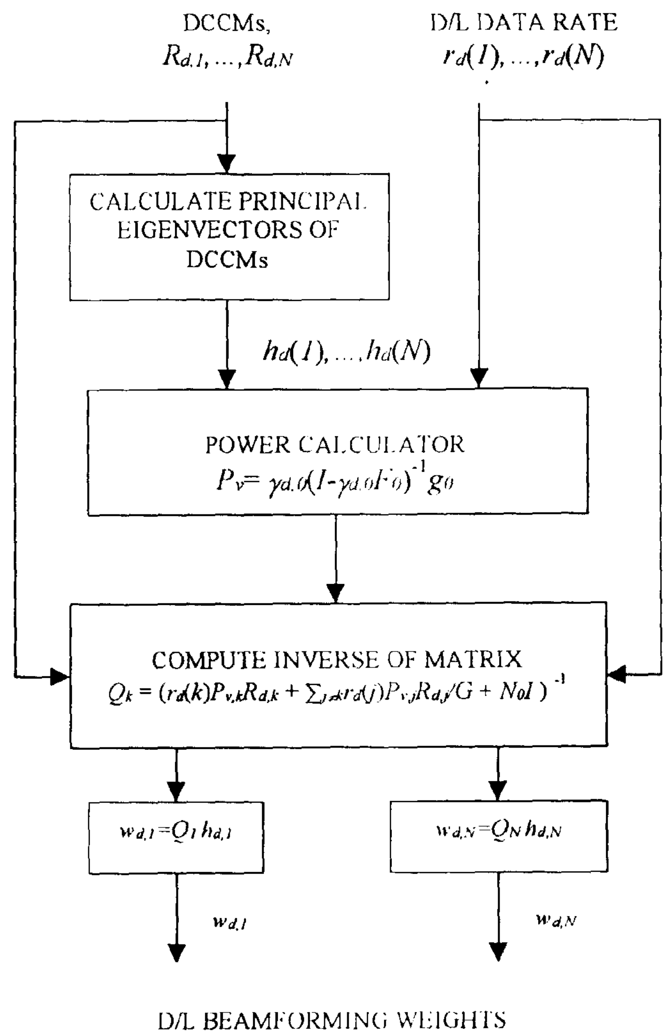

DCCMs,
$R_{d,1}, \ldots, R_{d,N}$
D/L DATA RATE
$r_d(1), \ldots, r_d(N)$
CALCULATE PRINCIPAL EIGENVECTORS OF DCCMs
$h_d(1), \ldots, h_d(N)$
POWER CALCULATOR
$P_v = \gamma_{d,0}(I - \gamma_{d,0}F_0)^{-1} g_0$
COMPUTE INVERSE OF MATRIX
$Q_k = (r_d(k)P_{v,k}R_{d,k} + \sum_{j \neq k} r_d(j)P_{v,j}R_{d,j}/G + N_0 I)^{-1}$
$w_{d,1} = Q_1 h_{d,1}$
$w_{d,N} = Q_N h_{d,N}$
$w_{d,1}$
$w_{d,N}$
D/L BEAMFORMING WEIGHTS

FIG. 6

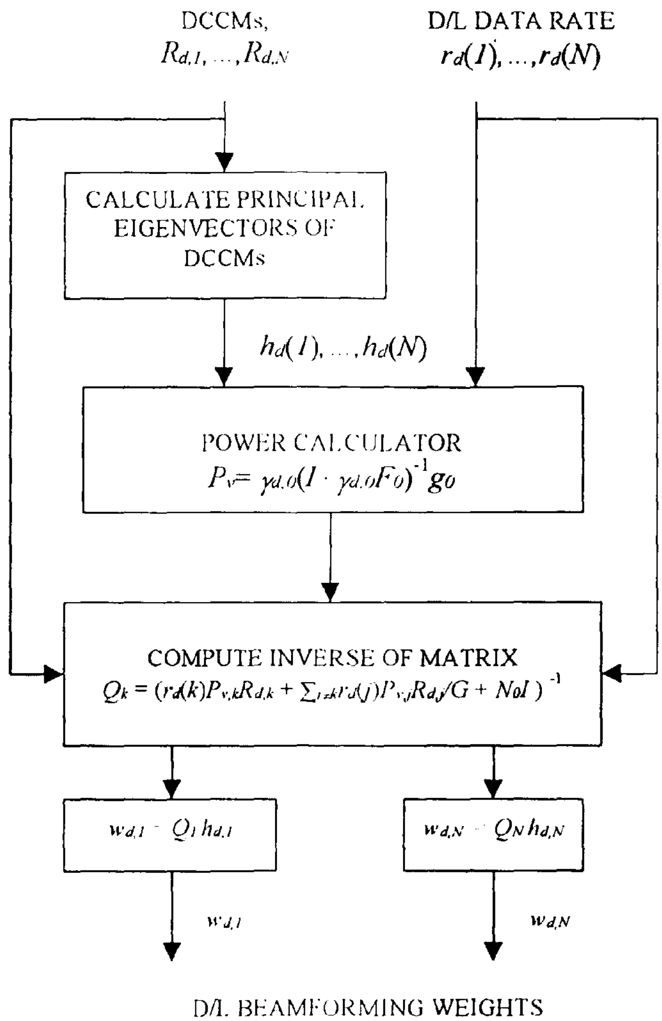

DCCMs,
$R_{d,1}, \ldots, R_{d,N}$
D/L DATA RATE
$r_d(1), \ldots, r_d(N)$
CALCULATE PRINCIPAL EIGENVECTORS OF DCCMs
$h_d(1), \ldots, h_d(N)$
POWER CALCULATOR
$P_v = \gamma_{d,0}(I - \gamma_{d,0}F_0)^{-1}g_0$
COMPUTE INVERSE OF MATRIX
$Q_k = (r_d(k)P_{v,k}R_{d,k} + \sum_{j \neq k} r_d(j)P_{v,j}R_{d,j}/G + N_0 I)^{-1}$
$w_{d,1} = Q_1 h_{d,1}$
$w_{d,N} = Q_N h_{d,N}$
$w_{d,1}$
$w_{d,N}$
D/L BEAMFORMING WEIGHTS

FIG. 7

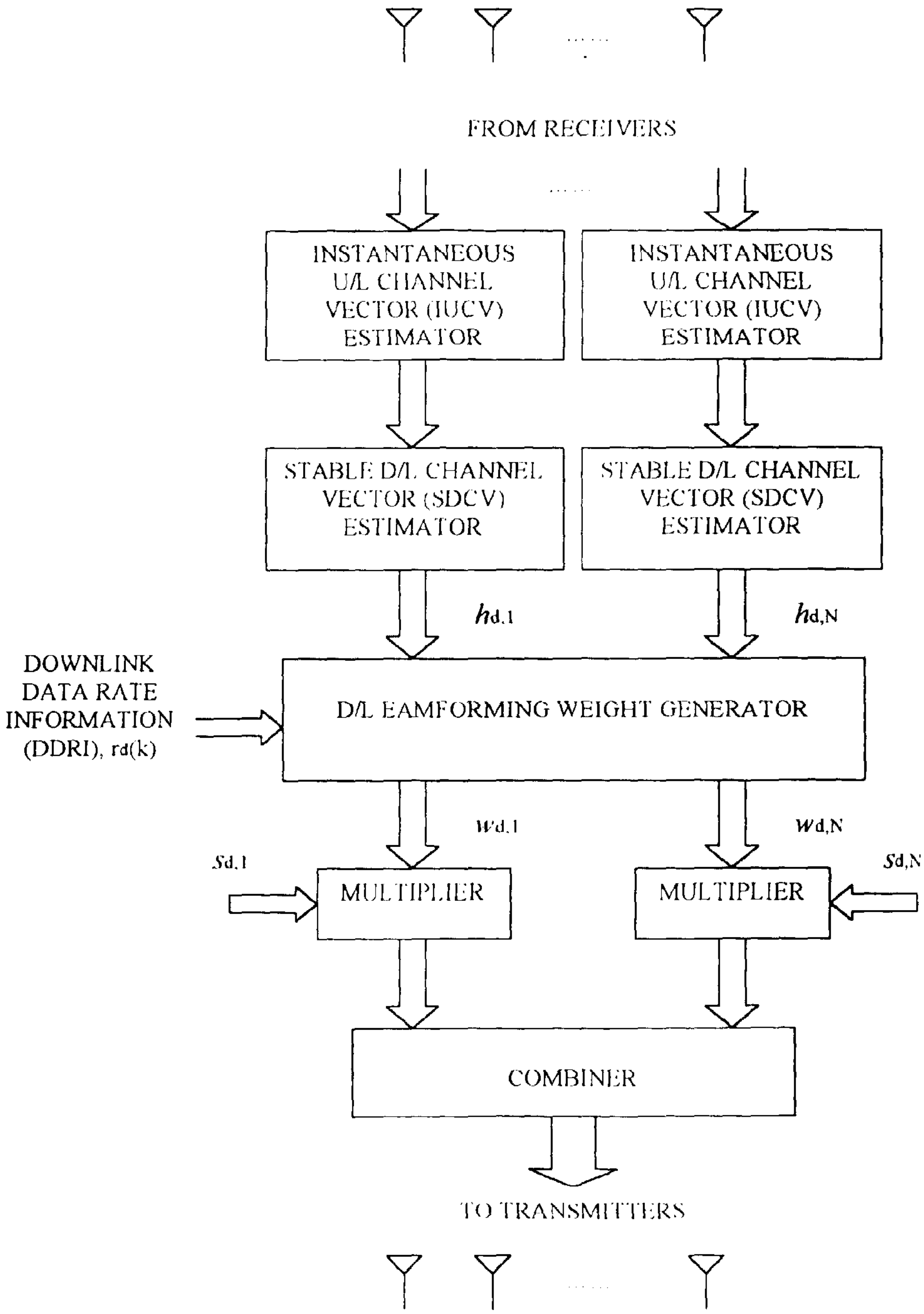

FROM RECEIVERS
INSTANTANEOUS U/L CHANNEL VECTOR (IUCV) ESTIMATOR
INSTANTANEOUS U/L CHANNEL VECTOR (IUCV) ESTIMATOR
STABLE D/L CHANNEL VECTOR (SDCV) ESTIMATOR
STABLE D/L CHANNEL VECTOR (SDCV) ESTIMATOR
$h_{d,1}$
$h_{d,N}$
DOWNLINK DATA RATE INFORMATION (DDRI), $r_d(k)$
D/L EAMFORMING WEIGHT GENERATOR
$w_{d,1}$
$w_{d,N}$
$s_{d,1}$
$s_{d,N}$
MULTIPLIER
MULTIPLIER
COMBINER
TO TRANSMITTERS

FIG.8

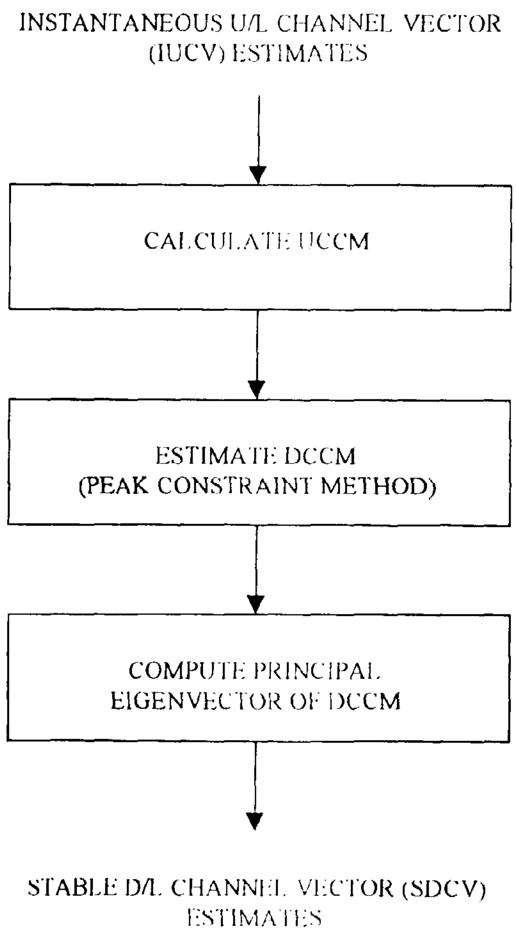

INSTANTANEOUS U/L CHANNEL VECTOR (IUCV) ESTIMATES
CALCULATE UCCM
ESTIMATE DCCM
(PEAK CONSTRAINT METHOD)
COMPUTE PRINCIPAL EIGENVECTOR OF DCCM
STABLE D/L CHANNEL VECTOR (SDCV) ESTIMATES

FIG. 9

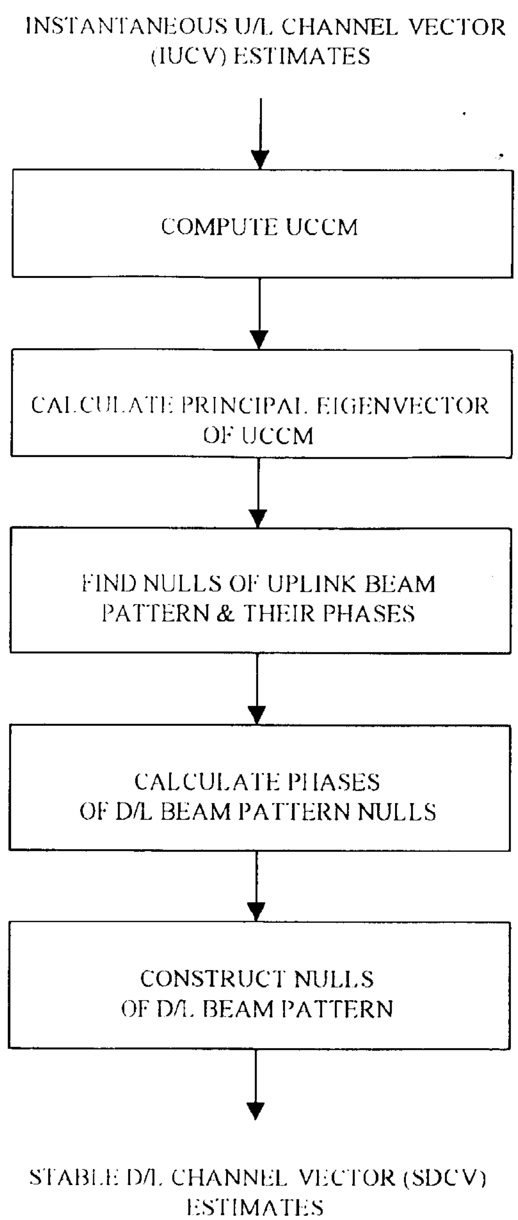

INSTANTANEOUS U/L CHANNEL VECTOR (IUCV) ESTIMATES
COMPUTE UCCM
CALCULATE PRINCIPAL EIGENVECTOR OF UCCM
FIND NULLS OF UPLINK BEAM PATTERN & THEIR PHASES
CALCULATE PHASES OF D/L BEAM PATTERN NULLS
CONSTRUCT NULLS OF D/L BEAM PATTERN
STABLE D/L CHANNEL VECTOR (SDCV) ESTIMATES

FIG. 10

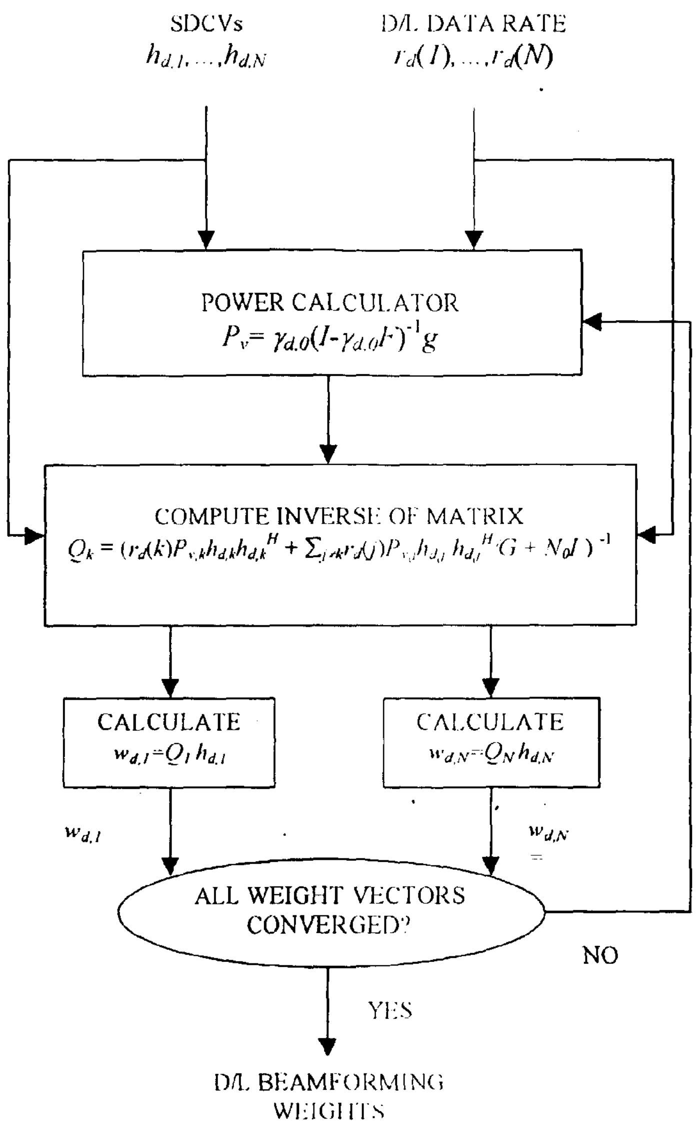

SDCVs
$h_{d,1}, \ldots, h_{d,N}$
D/L DATA RATE
$r_d(1), \ldots, r_d(N)$
POWER CALCULATOR
$P_v = \gamma_{d,0}(I - \gamma_{d,0}F)^{-1}g$
COMPUTE INVERSE OF MATRIX
$Q_k = (r_d(k)P_{v,k}h_{d,k}h_{d,k}^H + \sum_{j \neq k} r_d(j)P_{v,j}h_{d,j}h_{d,j}^H/G + N_0 I)^{-1}$
CALCULATE
$w_{d,1} = Q_1 h_{d,1}$
CALCULATE
$w_{d,N} = Q_N h_{d,N}$
$w_{d,1}$
$w_{d,N}$
ALL WEIGHT VECTORS CONVERGED?
NO
YES
D/L BEAMFORMING WEIGHTS

FIG.11

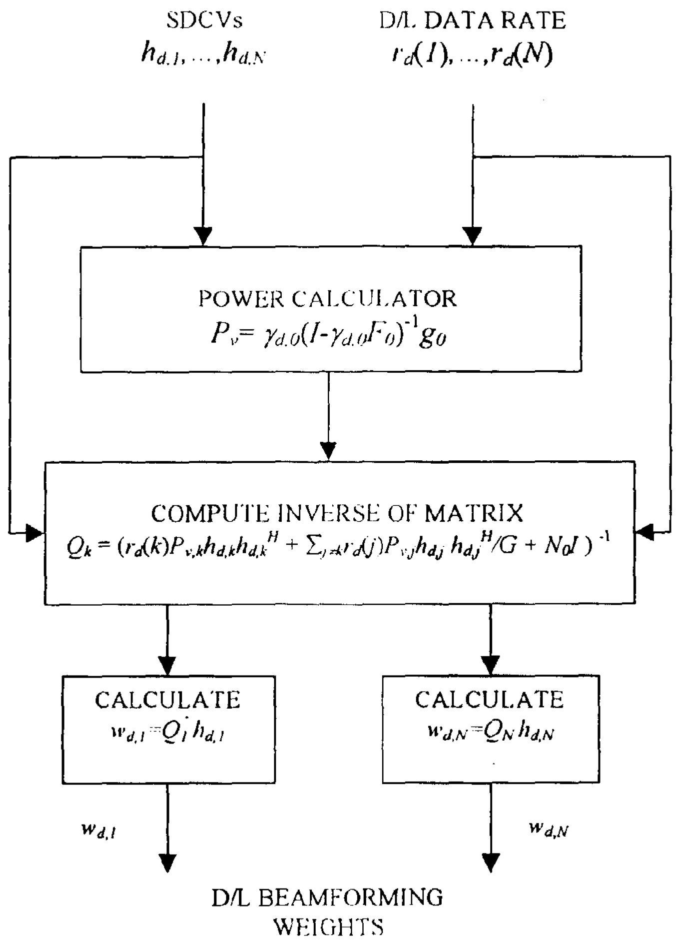

SDCVs
$h_{d,1},\ldots,h_{d,N}$
D/L DATA RATE
$r_d(1),\ldots,r_d(N)$
POWER CALCULATOR
$P_v = \gamma_{d,0}(I-\gamma_{d,0}F_0)^{-1}g_0$
COMPUTE INVERSE OF MATRIX
$Q_k = (r_d(k)P_{v,k}h_{d,k}h_{d,k}^H + \sum_{j\neq k} r_d(j)P_{v,j}h_{d,j}\,h_{d,j}^H/G + N_0I\,)^{-1}$
CALCULATE
$w_{d,1}=Q_1 h_{d,1}$
CALCULATE
$w_{d,N}=Q_N h_{d,N}$
$w_{d,1}$
$w_{d,N}$
D/L BEAMFORMING
WEIGHTS

FIG. 12

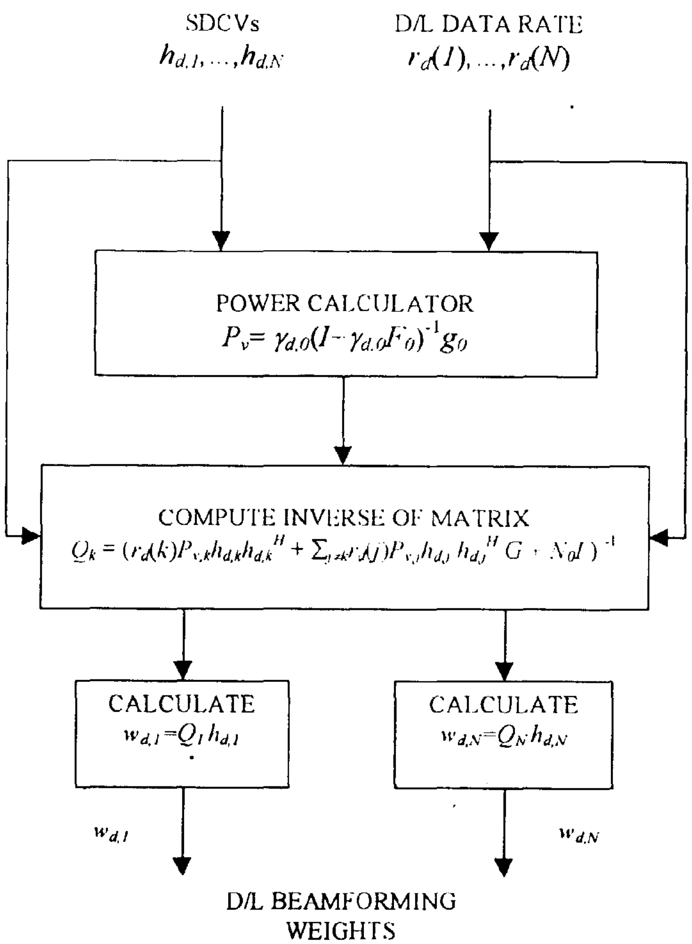

SDCVs
$h_{d,1},\ldots,h_{d,N}$
D/L DATA RATE
$r_d(1),\ldots,r_d(N)$
POWER CALCULATOR
$P_v = \gamma_{d,0}(I - \gamma_{d,0}F_0)^{-1}g_0$
COMPUTE INVERSE OF MATRIX
$Q_k = (r_d(k)P_{v,k}h_{d,k}h_{d,k}^H + \sum_{j\neq k} r_d(j)P_{v,j}h_{d,j}\,h_{d,j}^H G + N_0 I)^{-1}$
CALCULATE
$w_{d,1} = Q_1 h_{d,1}$
CALCULATE
$w_{d,N} = Q_N h_{d,N}$
$w_{d,1}$
$w_{d,N}$
D/L BEAMFORMING
WEIGHTS

FIG. 13

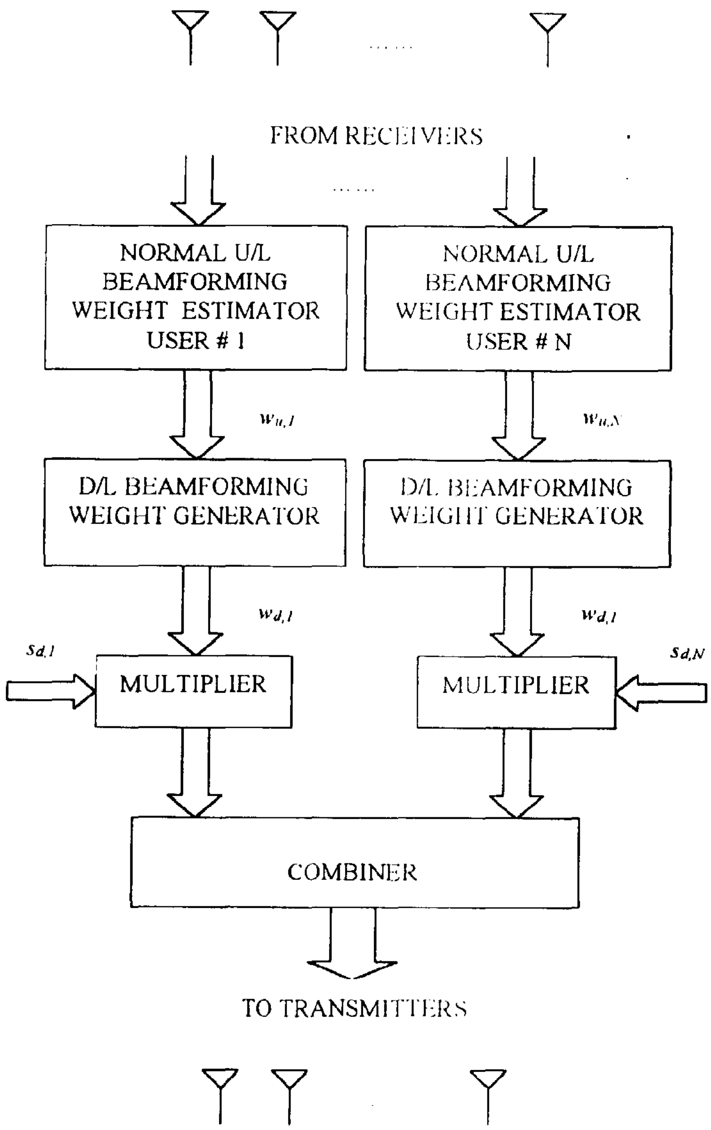

FROM RECEIVERS
NORMAL U/L BEAMFORMING WEIGHT ESTIMATOR USER # 1
NORMAL U/L BEAMFORMING WEIGHT ESTIMATOR USER # N
$w_{u,1}$
$w_{u,N}$
D/L BEAMFORMING WEIGHT GENERATOR
D/L BEAMFORMING WEIGHT GENERATOR
$w_{d,1}$
$w_{d,1}$
$s_{d,1}$
$s_{d,N}$
MULTIPLIER
MULTIPLIER
COMBINER
TO TRANSMITTERS

FIG.14

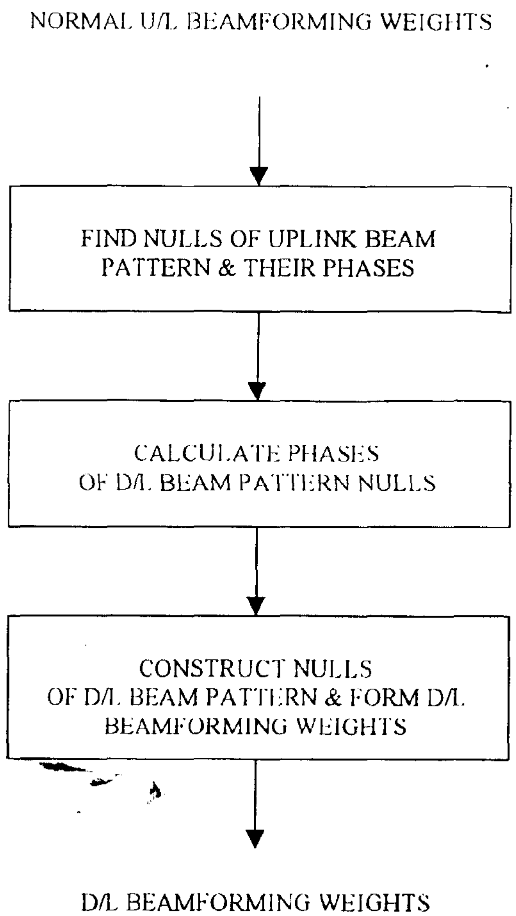

NORMAL U/L BEAMFORMING WEIGHTS
FIND NULLS OF UPLINK BEAM PATTERN & THEIR PHASES
CALCULATE PHASES OF D/L BEAM PATTERN NULLS
CONSTRUCT NULLS OF D/L BEAM PATTERN & FORM D/L BEAMFORMING WEIGHTS
D/L BEAMFORMING WEIGHTS

Fig. 15

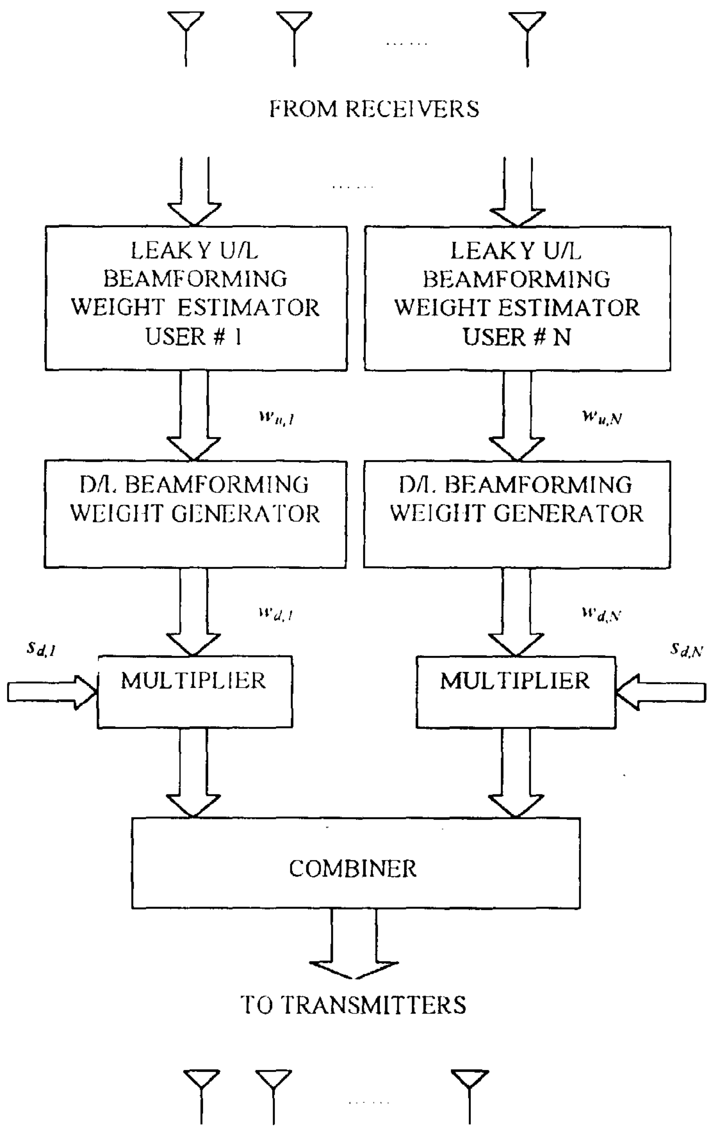

FROM RECEIVERS
LEAKY U/L BEAMFORMING WEIGHT ESTIMATOR USER # 1
LEAKY U/L BEAMFORMING WEIGHT ESTIMATOR USER # N
$w_{u,1}$
$w_{u,N}$
D/L BEAMFORMING WEIGHT GENERATOR
D/L BEAMFORMING WEIGHT GENERATOR
$w_{d,1}$
$w_{d,N}$
$s_{d,1}$
$s_{d,N}$
MULTIPLIER
MULTIPLIER
COMBINER
TO TRANSMITTERS

FIG. 16

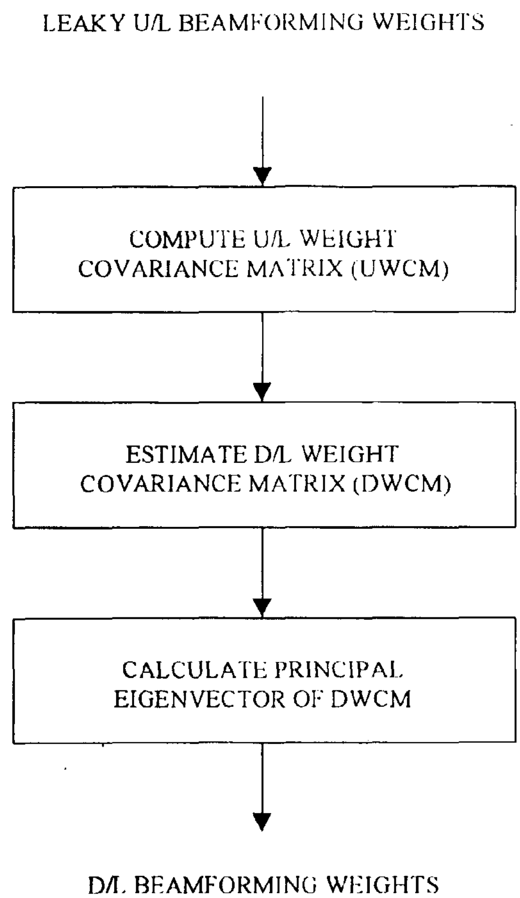

LEAKY U/L BEAMFORMING WEIGHTS
COMPUTE U/L WEIGHT COVARIANCE MATRIX (UWCM)
ESTIMATE D/L WEIGHT COVARIANCE MATRIX (DWCM)
CALCULATE PRINCIPAL EIGENVECTOR OF DWCM
D/L BEAMFORMING WEIGHTS

FIG. 17

European Patent Office

EUROPEAN SEARCH REPORT

Application Number
EP 00 11 8876

DOCUMENTS CONSIDERED TO BE RELEVANT

Category	Citation of document with indication, where appropriate, of relevant passages	Relevant to claim	CLASSIFICATION OF THE APPLICATION (Int.Cl.7)
E	WO 00 59072 A (BONEK ERNST ;HUGL KLAUS (AT); INNOVATIONSAGENTUR GES M B H (AT)) 5 October 2000 (2000-10-05) * page 7, line 22 - page 12, line 7 * ---	1-42	H04B7/08
A	WO 98 27669 A (CWILL TELECOMMUNICATIONS INC) 25 June 1998 (1998-06-25) * page 8, line 9 - page 21, line 3 * ---	1-42	
A	US 5 634 199 A (PAULRAJ AROGYASWAMI ET AL) 27 May 1997 (1997-05-27) * the whole document * ---	1-42	
A	US 5 886 988 A (OTTERSTEN BJOERN E ET AL) 23 March 1999 (1999-03-23) * the whole document * ---	1-42	
A	US 5 631 898 A (DENT PAUL W) 20 May 1997 (1997-05-20) * the whole document * ---	1-42	TECHNICAL FIELDS SEARCHED (Int.Cl.7)
A	US 5 572 216 A (WEINBERG AARON ET AL) 5 November 1996 (1996-11-05) * the whole document * -----	1-42	H04B

The present search report has been drawn up for all claims

Place of search	Date of completion of the search	Examiner
MUNICH	19 January 2001	Villafuerte Abrego

CATEGORY OF CITED DOCUMENTS

X : particularly relevant if taken alone
Y : particularly relevant if combined with another document of the same category
A : technological background
O : non-written disclosure
P : intermediate document

T : theory or principle underlying the invention
E : earlier patent document, but published on, or after the filing date
D : document cited in the application
L : document cited for other reasons

& : member of the same patent family, corresponding document

ANNEX TO THE EUROPEAN SEARCH REPORT ON EUROPEAN PATENT APPLICATION NO. EP 00 11 8876

This annex lists the patent family membersrelating to the patent documents cited in the above-mentioned European search report.
The members are as contained in the European Patent Office EDP file on
The European Patent Office is in no way liable for these particulars which are merely given for the purpose of information.

19-01-2001

Patent document cited in search report	Publication date	Patent family member(s)	Publication date
WO 0059072 A	05-10-2000	AT 56599 A	15-10-2000
WO 9827669 A	25-06-1998	US 6122260 A	19-09-2000
		AU 5524398 A	15-07-1998
		EP 0953235 A	03-11-1999
US 5634199 A	27-05-1997	US 5471647 A	28-11-1995
US 5886988 A	23-03-1999	US 5828658 A	27-10-1998
		AU 5714898 A	31-07-1998
		BR 9714121 A	29-02-2000
		EP 0956716 A	17-11-1999
		WO 9830047 A	09-07-1998
		AU 5086498 A	15-05-1998
		BR 9712643 A	26-10-1999
		CN 1234947 A	10-11-1999
		EP 0932986 A	04-08-1999
		WO 9818272 A	30-04-1998
US 5631898 A	20-05-1997	US 5619503 A	08-04-1997
		AU 1524595 A	01-08-1995
		BR 9505640 A	09-01-1996
		CA 2157182 A	13-07-1995
		CN 1124075 A	05-06-1996
		EP 0688482 A	27-12-1995
		FI 954210 A	19-10-1995
		JP 8510609 T	05-11-1996
		US 6148041 A	14-11-2000
		WO 9519078 A	13-07-1995
		US 6088593 A	11-07-2000
		US 5790606 A	04-08-1998
		US 5708971 A	13-01-1998
		US 5555257 A	10-09-1996
		US 5594941 A	14-01-1997
		US 5848060 A	08-12-1998
		US 5937015 A	10-08-1999
		US 5907809 A	25-05-1999
		US 5812947 A	22-09-1998
		US 5940742 A	17-08-1999
US 5572216 A	05-11-1996	AU 1289395 A	06-06-1995
		WO 9514335 A	26-05-1995
		US 6020845 A	01-02-2000

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(71) Applicant: **LUCENT TECHNOLOGIES INC.**
Murray Hill, New Jersey 07974-0636 (US)

(72) Inventors:
- **Foschini, Gerard Joseph**
  Sayerville, New Jersey 08879 (US)
- **Lozano, Angel**
  Holmdel, New Jersey 07733 (US)
- **Rashid-Farrokhi, Farrokh**
  Manalapan, New Jersey 07726 (US)
- **Valenzuela, Reinaldo A.**
  Holmdel, New Jersey 07733 (US)

(74) Representative:
Watts, Christopher Malcolm Kelway, Dr. et al
Lucent Technologies (UK) Ltd,
5 Mornington Road
Woodford Green Essex, IG8 0TU (GB)

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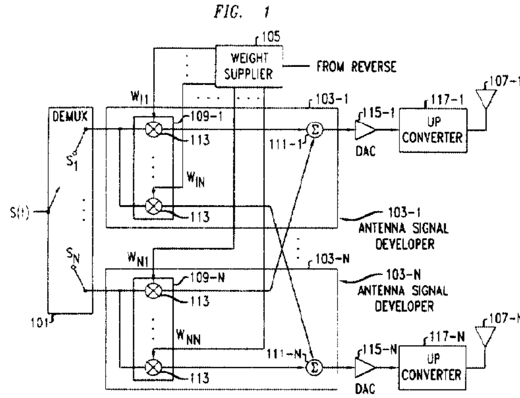
A request for correction of claims 16 to 18 has been filed pursuant to Rule 88 EPC. A decision on the request will be taken during the proceedings before the Examining Division (Guidelines for Examination in the EPO, A-V, 3.).

(54) **Space-time processing for multiple-input multiple-output wireless communication systems**

(57) In a MIMO system the signals transmitted from the various antennas are processed so as to improve the ability of the receiver to extract them from the received signal even in the face of some correlation. More specifically the number of bit streams that is transmitted simultaneously is adjusted, e.g., reduced, depending on the level of correlation, while multiple versions of each bit stream, variously weighted, are transmitted simultaneously. The variously weighted versions are combined to produced one combined weighted signal. The receiver processes the received signals in the same manner as it would have had all the signals reaching the receive antennas been uncorrelated. The weight vectors may be determined by the forward channel transmitter using the channel properties of the forward link which are made known to the transmitter of the forward link by being transmitted from the receiver of the forward link by the transmitter of the reverse link or the weight vectors may be determined by the forward channel transmitter using the channel properties of the forward link and the determined weight vectors are made known to the transmitter of the forward link by being transmitted from the receiver of the forward link by the transmitter of the reverse link. The channel properties used to determine the weight vectors may include the channel response from the transmitter to the receiver and the covariance matrix of noise and interference measured at the receiver.

FIG. 1



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Description

Technical Field

[0001] This invention relates to the art of wireless communications, and more particularly, to wireless communication systems using multiple antennas at the transmitter and multiple antennas at the receivers, so called multiple-input, multiple-output (MIMO) systems.

Background of the Invention

[0002] It is well known in the art that multiple-input, multiple-output (MIMO) systems can achieve dramatically improved capacity as compared to single antenna, i.e., single antenna to single antenna or multiple antenna to single antenna, systems. However, to achieve this improvement, it is preferable that there be a rich scattering environments, so that the various signals reaching the multiple receive antennas be largely uncorrelated. If the signals have some degree of correlation, and such correlation is ignored, performance degrades and capacity is reduced.

Summary of the Invention

[0003] We have invented a way of developing signals in a MIMO system such that even in the face of some correlation so as to obtain the most performance and capacity that can be achieved with a channel of that level of correlation. In accordance with the principles of the invention, the signals transmitted from the various antennas are processed so as to improve the ability of the receiver to extract them from the received signal. More specifically the number of bit streams that is transmitted simultaneously is adjusted, e.g., reduced, depending on the level of correlation, while multiple versions of each bit stream, variously weighted, are transmitted simultaneously. The variously weighted versions are combined to produced one combined weighted signal, a so-called "transmit vector", for each antenna. The receiver processes the received signals in the same manner as it would have had all the signals reaching the receive antennas been uncorrelated.

[0004] In one embodiment of the invention, the weight vectors are determined by the forward channel transmitter using the channel properties of the forward link which are made known to the transmitter of the forward link by being transmitted from the receiver of the forward link by the transmitter of the reverse link. In another embodiment of the invention the weight vectors are determined by the forward channel receiver using the channel properties of the forward link and the determined weight vectors are made known to the transmitter of the forward link by being transmitted from the receiver of the forward link by the transmitter of the reverse link.

[0005] The channel properties used to determine the weight vectors may include the channel response from the transmitter to the receiver and the covariance matrix of noise and interference measured at the receiver.

Brief Description of the Drawing

[0006] In the drawing:

FIG. 1 shows an exemplary portion of a transmitter for developing signals to transmit in a MIMO system such that even in the face of some correlation the most performance and capacity that can be achieved with a channel of that level of correlation is obtained, in accordance with the principles of the invention;

FIG. 2 shows an exemplary portion of a receiver for a MIMO system arranged in accordance with the principles of the invention; and

FIG. 3 shows an exemplary process, in flow chart form, for developing signals to transmit in a MIMO system such that even in the face of some correlation the most performance and capacity that can be achieved with a channel of that level of correlation is obtained, in accordance with the principles of the invention;

FIG. 4 shows another exemplary process, in flow chart form, for developing signals to transmit in a MIMO system such that even in the face of some correlation the most performance and capacity that can be achieved with a channel of that level of correlation is obtained, in accordance with the principles of the invention.

Detailed Description

[0007] The following merely illustrates the principles of the invention. It will thus be appreciated that those skilled in the art will be able to devise various arrangements which, although not explicitly described or shown herein, embody the principles of the invention and are included within its spirit and scope. Furthermore, all examples and conditional language recited herein are principally intended expressly to be only for pedagogical purposes to aid the reader in understanding the principles of the invention and the concepts contributed by the inventor(s) to furthering the art, and are to be construed as being without limitation to such specifically recited examples and conditions. Moreover, all statements herein reciting principles, aspects, and embodiments of the invention, as well as specific examples thereof, are intended to encompass both structural and functional equivalents thereof. Additionally, it is intended that such equivalents include both currently known equivalents as well as equivalents developed in the future, i.e., any elements developed that perform the same function, regardless of structure.

[0008] Thus, for example, it will be appreciated by those skilled in the art that the block diagrams herein represent conceptual views of illustrative circuitry em-

bodying the principles of the invention. Similarly, it will be appreciated that any flow charts, flow diagrams, state transition diagrams, pseudocode, and the like represent various processes which may be substantially represented in computer readable medium and so executed by a computer or processor, whether or not such computer or processor is explicitly shown.

[0009] The functions of the various elements shown in the FIGs., including functional blocks labeled as "processors" may be provided through the use of dedicated hardware as well as hardware capable of executing software in association with appropriate software. When provided by a processor, the functions may be provided by a single dedicated processor, by a single shared processor, or by a plurality of individual processors, some of which may be shared. Moreover, explicit use of the term "processor" or "controller" should not be construed to refer exclusively to hardware capable of executing software, and may implicitly include, without limitation, digital signal processor (DSP) hardware, read-only memory (ROM) for storing software, random access memory (RAM), and non-volatile storage. Other hardware, conventional and/or custom, may also be included. Similarly, any switches shown in the FIGS. are conceptual only. Their function may be carried out through the operation of program logic, through dedicated logic, through the interaction of program control and dedicated logic, or even manually, the particular technique being selectable by the implementor as more specifically understood from the context.

[0010] In the claims hereof any element expressed as a means for performing a specified function is intended to encompass any way of performing that function including, for example, a) a combination of circuit elements which performs that function or b) software in any form, including, therefore, firmware, microcode or the like, combined with appropriate circuitry for executing that software to perform the function. The invention as defined by such claims resides in the fact that the functionalities provided by the various recited means are combined and brought together in the manner which the claims call for. Applicant thus regards any means which can provide those functionalities as equivalent as those shown herein.

[0011] FIG. 1 shows an exemplary portion of a transmitter for developing signals to transmit in a MIMO system having a transmitter with N transmit antennas transmitting over a forward channel to a receiver having L receiver antennas and a reverse channel for communicating from said receiver to said transmitter, such that even in the face of some correlation the most performance and capacity that can be achieved with a channel of that level of correlation is obtained, in accordance with the principles of the invention. Shown in FIG. 1 are a) demultiplexer (demux) 101; b) antenna signal developers 103, including antenna signal developers 103-1 through 103-N; c) weight supplier 105; d) N antennas 107, including antennas 107-1 through 107-N; e) digital-to-analog converters (DAC) 115, including 115-1 through 115-N; and f) upconverters 117, including upconverters 117-1 through 117-N.

[0012] Demultiplexer 101 takes a data stream as an input and supplies as an output data substreams by supplying various bits from the input data stream to each of the data substreams. One data substream may be supplied by demultiplexer 101 to one of N outputs. However, when the number of uncorrelated signals that can be transmitted is reduced, the number of bit streams that are transmitted simultaneously is reduced to match the number of uncorrelated signals that can be transmitted. In such a case, the particular outputs utilized is at the discretion of the implementor. For example, only the first Y outputs, where Y is the number of uncorrelated signals that can be transmitted, are employed.

[0013] Each data substream is supplied to a corresponding one of antenna signal developers 103. Each one of antenna signal developers 103 includes one of weight blocks 109-1 through 109-N and one of adders 111-1 through 111-N. Within each of antenna signal developers 103 the data substream is supplied to each of multipliers 113 within the one of weight blocks 109 therein.

[0014] Weight supplier 105 supplies weight values to each of multipliers 113. In one embodiment of the invention weight supplier 105 actually develops the weight values in response to information received via the reverse channel from the receiver (not shown). In another embodiment of the invention the weight values are developed in the receiver, then supplied via the reverse channel to the transmitter, in which they are stored in weight supplier 105 until such time as they are required. A process for developing the weights in accordance with an aspect of the invention will be described hereinbelow.

[0015] Each of multipliers 113 multiplies the substream it receives by the weight it receives. The resulting product is supplied to a respective one of adders 111. More specifically, the product supplied by the Rth multiplier of each weight block 109, where R is from 1 to N, is supplied to the Rth one of adders 111. For those multipliers that are not supplied with a substream, their output is insured to be zero (0), by any technique desired by the implementor.

[0016] Each of adders 111 adds the signals input to it and supplies the resulting sum as an output to its associated respective one of DACs 115. Each of DACs 115 takes the digital signal it receives from one of adders 111 and converts it to an analog baseband signal. The analog baseband signal produced by each of DACs 115 is supplied to a respective one of upconverters 117, which upconverts the baseband analog signal to a radio frequency signal. The radio frequency signals produced by upconverters 117 are supplied to respective ones of antennas 107 for broadcast to a receiver.

[0017] FIG. 2 shows an exemplary portion of a receiver for a MIMO system arranged in accordance with the principles of the invention. FIG. 2 shows a) L antennas

201, including antennas 201-1 through 201-L; b) downconverters 203, including downconverters 203-1 through 203-L; c) analog-to-digital converters (ADCs) 205, including analog-to-digital converters 205-1 through 205-L; d) estimate interference covariance matrix and channel response unit 207; e) optional weight calculator 209; and f) optional switch 211.

[0018] Each of antennas 201 receives radio signals and supplies an electrical version thereof to its respective, associated one of downconverters 203. Each of downconverters 203 downconverts the signal it receives to baseband, and supplies the resulting baseband signal to its associated one of ADCs 205. Each of ADCs 205 converts the baseband analog signal it received to a digital representation and supplies the digital representation to estimate interference covariance matrix and channel response unit 207.

[0019] Estimate interference covariance matrix and channel response unit 207 develops an estimate of the interference covariance matrix and an estimate of the forward matrix channel response in the conventional manner. Note that matrices are required because there are multiple transmit antennas and multiple receive antennas.

[0020] The estimate of the interference covariance matrix and an estimate of the forward matrix channel response are supplied either to optional weight calculator 209 or they are supplied for via the reverse channel to the transmitter (FIG. 1). If the estimate of the interference covariance matrix and an estimate of the forward matrix channel response is supplied to weight calculator 209, weight calculator determines the weight values that are to be used, in accordance with an aspect of the invention and as described hereinbelow, and supplies the resulting weight values to the transmitter (FIG. 1) via the reverse channel.

[0021] FIG. 3 shows an exemplary process, in flow chart form, for developing signals to transmit in a MIMO system having a transmitter with N transmit antennas transmitting over a forward channel to a receiver having L receiver antennas and a reverse channel for communicating from said receiver to said transmitter, such that even in the face of some correlation the most performance and capacity that can be achieved with a channel of that level of correlation is obtained, in accordance with the principles of the invention. The process of FIG. 3 may be employed in an embodiment of the invention that uses the hardware of FIGs. 1 and 2, with switch 211 being connected to estimate interference covariance matrix and channel response unit 207 and with a communication protocol as follows. First it is necessary to determine the length of time during which the channel characteristics are stable. This is typically performed at the system engineering phase of developing the system, using measurements of the environment into which the system is to be deployed, as is well known by those of ordinary skill in the art. Once the length of time for which the channel characteristics are stable is known, that time is considered as a frame, and the frame is divided into time slots. Each frame has a preamble, which may occupy one or more of the time slots. The frames, and accordingly the time slots, are repeating in nature.

[0022] The process of FIG. 3 is entered in step 301 at the beginning of each frame. Next, in step 303, the interference covariance matrix K^N and channel response H at the receiver are determined, e.g., in the receiver of the forward link, such as in interference covariance matrix and channel response unit 207 (FIG. 2). Thereafter, in step 305, (FIG. 3) interference covariance matrix K^N and channel response matrix H are supplied by the receiver of the forward link to the transmitter of forward link, e.g., via the reverse channel.

[0023] In step 307 weights $w_i = [w_{i1},...,w_{iN}]$ are calculated, e.g., by weight supplier 105 (FIG. 1), where *i* is an integer ranging from 1 to N. More specifically, the weights are calculated as follows. First the matrix equation $H^\dagger(K^N)H = U^\dagger\Lambda^2 U$ is solved, where:

a) H is the channel response matrix;
b) $H^\dagger$ is the conjugate transpose of channel response matrix H, † being the well known symbol for conjugate transpose;
c) K^N is the interference covariance matrix;
d) U is a unitary matrix, each column of which is an eigenvector of $H^\dagger (K^N)H$;
e) Λ is a diagonal matrix defined as $\Lambda = diag(\lambda^1,...,\lambda^M)$, where $\lambda^1,...,\lambda^M$ are each eignevalues of $H^\dagger (K^N)$ H, M being the maximum number of nonzero eigenvalues, which corresponds to the number of substreams that actually can be used; and
f) $U^\dagger$ is the conjugate transpose of matrix U.

[0024] Then well known, so-called "waterfilling" is performed on the eigenvalues λ by solving the simultaneous equations $\tilde{\lambda}^k = (\nu - \frac{1}{(\lambda^k)^2})^+$ and $\sum_k \tilde{\lambda}^k = P$, for ν, where:

k is an integer index that ranges from 1 to M;
P is the transmitted power;
+ is an operator that returns zero (0) when its argument is negative, and returns the argument itself when it is positive; and
each $\tilde{\lambda}$ is an intermediate variable representative of a power for each weight vector.

[0025] A new matrix Φ is defined as $\Phi = U^\dagger diag(\tilde{\lambda}^1,...,\tilde{\lambda}^M)U$, where *diag* indicates that the various $\tilde{\lambda}$ are arranged as the elements of the main diagonal of the matrix, all other entries being zero (0). Each column of matrix Φ is used as a normalized, i.e., based on unit power, weight vector as indicated by $\Phi = [z_1,...,z_N]$ and the weight vectors are made up of individual weights *z*, $z_i = [z_{i1},..., z_{iN}]$. The weight vector $w_i = [w_{i1},..., w_{iN}]$ is then determined by unnormalizing, based on the power to be assigned to the weight vector, the various weights therein, being $\sqrt{\tilde{\lambda}^i} z_{ij}$, where *j* is an integer ranging from 1 to N.

[0026] In step 309 the input data stream, S(t) (FIG. 1), is divided into N substreams $S_1 \ldots S_N$, e.g., by demultiplexer 101. Each of the data streams is then multiplied by a respective one of weight vectors $w_{i1}, \ldots, w_{iN}$, in step 311 (FIG. 3). In other words, each bit of each of each particular data stream is multiplied by each of the weights in its respective weight vector to produce N weighted bits for each data stream.

[0027] In step 313 the weighted bits for each of the substreams is combined by each antenna adder, e.g., adders 111. In this regard, the weighted bit produced for each substream from the first weight is added at the adder of the first antenna, the weighted bit produced for each substream from the second weight is added at the adder of the second antenna, and so forth, as indicated in FIG. 1. As will be readily apparent from the foregoing, any substream greater in number than M will be zero, since M corresponds to the number of substreams that actually can be used. Such zero substreams do not contribute to the sum produced by adders 111.

[0028] The process then exits in step 315.

[0029] FIG. 4 shows another exemplary process, in flow chart form, for developing signals to transmit in a MIMO system having a transmitter with N transmit antennas transmitting over a forward channel to a receiver having L receiver antennas and a reverse channel for communicating from said receiver to said transmitter, such that even in the face of some correlation the most performance and capacity that can be achieved with a channel of that level of correlation is obtained, in accordance with the principles of the invention. The process of FIG. 4 may be employed in an embodiment of the invention that uses the hardware of FIGs. 1 and 2, with switch 211 being connected to weight calculator 209 and with a communication protocol as described in connection with FIG. 3. Note that for the process of FIG. 4, weight supplier 105 of FIG. 1 will not compute the various weights, but will instead merely store the weights received from weight calculator 209 and supply them to the various ones of multipliers 113 as is necessary.

[0030] The process of FIG. 4 is entered in step 401 at the beginning of each frame. Next, in step 403, the interference covariance matrix K^N and channel response H at the receiver are determined, e.g., in the receiver of the forward link, such as in interference covariance matrix and channel response unit 207 (FIG. 2). In step 405 weights $w_i = [w_{i1}, \ldots, w_{iN}]$ are calculated, e.g., by weight supplier 105 (FIG. 1). More specifically, the weights are calculated as follows.

[0031] First the matrix equation $H^\dagger(K^N)H = U^\dagger \Lambda^2 U$ is solved, where:

a) H is the channel response matrix;

b) $H^\dagger$ is the conjugate transpose of channel response matrix H, † being the well known symbol for conjugate transpose;

c) K^N is the interference covariance matrix;

d) U is a unitary matrix, each column of which is an eigenvector of $H^\dagger(K^N)H$;

e) Λ is a diagonal matrix defined as $\Lambda = diag(\lambda^1, \ldots, \lambda^M)$, where $\lambda^1, \ldots, \lambda^M$ are each eignevalues of $H^\dagger (K^N) H$, M being the maximum number of nonzero eigenvalues, which corresponds to the number of substreams that actually can be used; and

f) $U^\dagger$ is the conjugate transpose of matrix U.

[0032] Then well known, so-called "waterfilling" is performed on the eigenvalues λ by solving the simultaneous equations $\tilde{\lambda}^k = (\nu - \frac{1}{(\lambda^k)^2})^+$ and $\sum_k \tilde{\lambda}^k = P$, for ν, where:

k is an integer index that ranges from 1 to M;

P is the transmitted power;

+ is an operator that returns zero (0) when its argument is negative, and returns the argument itself when it is positive; and

each $\tilde{\lambda}$ is an intermediate variable representative of a power for each weight vector.

[0033] A new matrix Φ is defined as $\Phi = U^\dagger diag(\tilde{\lambda}^1, \ldots, \tilde{\lambda}^M)U$, where *diag* indicates that the various $\tilde{\lambda}$ are arranged as the elements of the main diagonal of the matrix, all other entries being zero (0). Each column of matrix Φ is used as a normalized, i.e., based on unit power, weight vector as indicated by $\Phi = [z_1, \ldots, z_N]$ and the weight vectors are made up of individual weights z, $z_i = [z_{i1}, \ldots, z_{iN}]$. The weight vector $w_i = [w_{i1}, \ldots, w_{iN}]$ is then determined by unnormalizing, based on the power to be assigned to the weight vector, the various weights therein being $\sqrt{\tilde{\lambda}^i} z_{ij}$, where j is an integer ranging from 1 to N.

[0034] Thereafter, in step 407, the determined weight values are supplied by the receiver of the forward link to the transmitter of forward link, e.g., via the reverse channel. The weights are stored in weight supplier 105 (FIG. 1)

[0035] In step 409 (FIG. 4) the input data stream, S(t) (FIG. 1), is divided into N substreams $S_1 \ldots S_N$, e.g., by demultiplexer 101. Each of the data streams is then multiplied by a respective one of weight vectors $w_{i1}, \ldots, w_{iN}$, in step 411 (FIG. 4), where i is an integer ranging from 1 to N. In other words, each bit of each of each particular data stream is multiplied by each of the weights in its respective weight vector to produce N weighted bits for each data stream.

[0036] In step 413 the weighted bits for each of the substreams is combined by each antenna adder, e.g., adders 111. In this regard, the weighted bit produced for each substream from the first weight is aided at the adder of the first antenna, the weighted bit produced for each substream from the second weight is added at the adder of the second antenna, and so forth, as indicated in FIG. 1. As will be readily apparent from the foregoing, any substream greater in number than M will be zero, since M corresponds to the number of substreams that actually can be used. Such zero substreams do not contribute to the sum produced by adders 111.

[0037] The process then exits in step 415.

[0038] In another embodiment of the invention, for use with so-called "time division duplex" (TDD) systems, which share a single channel for both the forward and reverse channels, the estimation of the channel response may be performed at either end of the wireless link. This is because since the forward and reverse channels share the same frequency channel, alternating between which is using the channel at any one time, then provided the time split between the forward and reverse channel is small, the channel response for the forward and reverse channels will be the same. Therefore, the receiver of the reverse channel will experience the same channel response as the receiver of the forward channel, and so the receiver of the reverse link can perform all the channel estimations that were previously performed by the receiver of the forward link. Likewise, the receiver of the forward channel will experience the same channel response as the receiver of the reverse channel, and so the receiver of the forward link can perform all the channel estimations that were previously performed by the receiver of the reverse link.

Claims

1. A method for transmitting signals in communications system having a transmitter with N transmit antennas transmitting over a forward channel to a receiver having L receiver antennas and a reverse channel for communicating from said receiver to said transmitter, in which there may exist correlation in the signals received by two or more of said L receive antennas, the method comprising the steps of:

    determining the number of independent signals that can be transmitted from said N transmit antennas to said L receive antennas;
    creating, from a data stream, a data substream to be transmitted for each of the number of independent signals that can be transmitted from said N transmit antennas to said L receive antennas;
    weighting each of said substreams with N weights, one weight for each of said N transmit antennas, to produce N weighted substreams per substream;
    combining one of said weighted substreams produced from each of said substreams for each of said transmit antennas to produce a transmit signal for each of said transmit antennas.

2. The method as defined in claim 1 further comprising the step of transmitting said transmit signal from a respective one of said antennas.

3. The method as defined in claim 1 further comprising the step of receiving said weights via said reverse channel.

4. The method as defined in claim 1 wherein said weights are determined by said transmitter as a function of channel information and interference covariance received from said receiver via said reverse channel.

5. The method as defined in claim 1 wherein said weights are determined by solving a matrix equation $H^{\dagger}(K^N)H = U^{\dagger}\Lambda^2 U$ where:

    H is a channel response matrix,
    $H^{\dagger}$ is a conjugate transpose of said channel response matrix H,
    K^N is the interference covariance matrix,
    U is a unitary matrix, each column of which is an eigenvector of $H^{\dagger}(K^N)H$,
    Λ is a diagonal matrix defined as $\Lambda = diag(\lambda^1,...,\lambda^M)$, where $\lambda^1,...,\lambda^M$ are each eignevalues of $H^{\dagger}(K^N)H$, M being the maximum number of nonzero eigenvalues, which corresponds to the number of said independent signals, and
    $U^{\dagger}$ is the conjugate transpose of matrix U;
    waterfilling said eigenvalues λ by solving the simultaneous equations $\tilde{\lambda}^k = (\nu - \frac{1}{(\lambda^k)^2})^+$ and $\sum_k \tilde{\lambda}^k = P$, for ν, where:
    k is an integer index that ranges from 1 to M,
    P is the transmitted power,
    + is an operator that returns zero (0) when its argument is negative, and returns the argument itself when it is positive, and
    each $\tilde{\lambda}$ is an intermediate variable representative of a power for each weight vector;
    defining matrix Φ as $\Phi = U^{\dagger}\, diag(\tilde{\lambda}^1,...,\tilde{\lambda}^M)U$, where *diag* indicates that the various $\tilde{\lambda}$ are arranged as the elements of the main diagonal of matrix Φ;
    wherein each column of matrix Φ is used as a normalized weight vector indicated by $\Phi = [z_1,...,z_N]$ and said normalized weight vectors are made up of individual normalized weights z, $z_i = [z_{i1},...,z_{iN}]$, where i is an integer ranging from 1 to N;
    developing an unnormalized weight vector $w_i = [w_{i1},...,w_{iN}]$, with each of said weights therein being $\sqrt{\tilde{\lambda}^i} z_{ij}$, where j is an integer ranging from 1 to N.

6. Apparatus for transmitting signals in communications system having a transmitter with N transmit antennas transmitting over a forward channel to a receiver having L receiver antennas and a reverse channel for communicating from said receiver to said transmitter, in which there may exist correlation in the signals received by two or more of said L receive antennas, the apparatus comprising:

means for determining the number of independent signals that can be transmitted from said N transmit antennas to said L receive antennas;

means for creating, from a data stream, a data substream to be transmitted for each of the number of independent signals that can be transmitted from said N transmit antennas to said L receive antennas;

means for weighting each of said substreams with N weights, one weight for each of said N transmit antennas, to produce N'weighted substreams per substream;

means for combining one of said weighted substreams produced from each of said substreams for each of said antennas to produce a transmit signal for each antenna.

7. Apparatus as defined in claim 6 wherein said transmitter comprises means for developing said weights.

8. Apparatus as defined in claim 6 wherein said transmitter comprises means for storing said weights.

9. Apparatus as defined in claim 6 wherein said receiver comprises means for developing said weights.

10. A transmitter for transmitting signals in communications system having a transmitter with N transmit antennas transmitting over a forward channel to a receiver having L receiver antennas and a reverse channel for communicating from said receiver to said transmitter, in which there may exist correlation in the signals received by two or more of said L receive antennas, the apparatus comprising:

a demultiplexor for creating, from a data stream, a data substream to be transmitted for each of the number of independent signals that can be transmitted from said N transmit antennas to said L receive antennas

multipliers for weighting each of said substreams with N weights, one weight for each of said N transmit antennas, to produce N weighted substreams per substream, each of said weights being a function of at least an estimate interference covariance matrix and an estimate of a forward matrix channel response between said transmitter and said receiver; and

adders for combining one of said weighted substreams produced from each of said substreams for each of said antennas to produce a transmit signal for each of said transmit antennas.

11. The transmitter as defined in claim 10 further comprising a digital to analog converter for converting each of said combined weighted substreams.

12. The transmitter as defined in claim 10 further comprising an upconverter for converting to radio frequencies each of said analog-converted combined weighted substreams.

13. The transmitter is defined in claim 10 wherein said weights are determined in said transmitter in response to said interference covariance matrix estimate and said estimate of the forward channel response received from said receiver over said reverse channel.

14. The transmitter as defined in claim 10 wherein said weights are determined in said receiver and are transmitted to said transmitter over said reverse channel.

15. The transmitter as defined in claim 10 wherein said weights are determined by solving a matrix equation $H^{\dagger}(K^N)H = U^{\dagger}\Lambda^2 U$ where:

H is a channel response matrix,

$H^{\dagger}$ is a conjugate transpose of said channel response matrix H,

K^N is the interference covariance matrix,

U is a unitary matrix, each column of which is an eigenvector of $H^{\dagger}(K^N)H$,

Λ is a diagonal matrix defined as $\Lambda = diag(\lambda^1,\ldots,\lambda^M)$, where $\lambda^1,\ldots,\lambda^M$ are each eignevalues of $H^{\dagger}(K^N)H$, M being the maximum number of nonzero eigenvalues, which corresponds to the number of said independent signals, and

$U^{\dagger}$ is the conjugate transpose of matrix U;

waterfilling said eigenvalues λ by solving the simultaneous equations $\tilde{\lambda}^k = (\nu - \frac{1}{(\lambda^k)^2})^+$ and $\sum_k \tilde{\lambda}^k = P$, for ν, where:

k is an integer index that ranges from 1 to M,

P is the transmitted power,

+ is an operator that returns zero (0) when its argument is negative, and returns the argument itself when it is positive, and

each $\tilde{\lambda}$ is an intermediate variable representative of a power for each weight vector;

defining matrix Φ as $\Phi = U^{\dagger} diag(\tilde{\lambda}^1,\ldots,\tilde{\lambda}^M)U$, where *diag* indicates that the various $\tilde{\lambda}$ are arranged as the elements of the main diagonal of matrix Φ;

wherein each column of matrix Φ is used as a normalized weight vector indicated by $\Phi = [z_1,\ldots,z_N]$ and said normalized weight vectors are made up of individual normalized weights z, $z_i = [z_{i1},\ldots, z_{iN}]$, where i is an integer ranging from 1 to N;

developing unnormalized weight vector $w_i = [w_{i1},\ldots,w_{iN}]$, with each of said weights therein being $\sqrt{\tilde{\lambda}^i} z_{ij}$, where j is an integer ranging from 1

to N.

16. The transmitter as defined in claim 10 wherein said transmitter and receiver communicate using time division multiplexing (TDD) and said weights are determined in said transmitter using an estimate of the forward channel response that is determined by a receiver of said reverse link for said transmitter.

L antennas;
L downconverters;
an estimator for determining an estimate of an interference covariance matrix for a forward channel being received by said receiver; and
a transmitter for a reverse channel for transmitting said estimate of an interference covariance matrix to a receiver for said reverse channel.

17. A receiver for use in a MIMO system, comprising:

L antennas;
L downconverters;
an estimator for determining an estimate of an interference covariance matrix for a forward channel being received by said receiver;

18. A receiver for use in a MIMO system, comprising:

an estimator for determining an estimate of a channel response for a forward channel being received by said receiver; and
a transmitter for a reverse channel for transmitting said estimate of an interference covariance matrix and said estimate of a channel response to a receiver for said reverse channel.

19. A receiver for use in a MIMO system, comprising:

an estimator for determining an estimate of an interference covariance matrix for a forward channel being received by said receiver;
an estimator for determining an estimate of a channel response for a forward channel being received by said receiver; and
a weight calculator for calculating weights for use by a transmitter of said forward channel to transmit data substreams to said receiver as a function of said estimate of an interference covariance matrix for a forward channel being received by said receiver and said estimate of a channel response for a forward channel being received by said receiver.

20. The invention as defined in claim 19 further including a transmitter for a reverse channel for transmitting said weights to a receiver for said reverse channel.

21. A receiver for use in a MIMO system, comprising:

L antennas;
L downconverters;
an estimator for determining an estimate of an interference covariance matrix for a forward channel being received by said receiver;
an estimator for determining an estimate of a channel response for a forward channel being received by said receiver; and
a weight calculator for calculating weights for use by a transmitter of said forward channel to transmit data substreams to said receiver, said weights being determined in said weight calculator by
solving a matrix equation $H^{\dagger}(K^N)H = U^{\dagger}\Lambda^2 U$ where:
H is a channel response matrix,
$H^{\dagger}$ is a conjugate transpose of said channel response matrix H,
K^N is the interference covariance matrix,
U is a unitary matrix, each column of which is an eigenvector of $H^{\dagger}(K^N)H$,
Λ is a diagonal matrix defined as $\Lambda = diag(\lambda^1,\ldots,\lambda^M)$, where $\lambda^1,\ldots,\lambda^M$ are each eignevalues of $H^{\dagger}(K^N)H$, M being the maximum number of nonzero eigenvalues, which corresponds to the number of said independent signals, and
$U^{\dagger}$ is the conjugate transpose of matrix U;
waterfilling said eigenvalues λ by solving the simultaneous equations $\tilde{\lambda}^k = (\nu - \frac{1}{(\lambda^k)^2})^+$ and $\sum_k \tilde{\lambda}^k = P$, for ν, where:
k is an integer index that ranges from 1 to M,
P is the transmitted power,
+ is an operator that returns zero (0) when its argument is negative, and returns the argument itself when it is positive, and
each $\tilde{\lambda}$ is an intermediate variable representative of a power for each weight vector;
defining matrix Φ as $\Phi = U^{\dagger}\, diag(\tilde{\lambda}^1,\ldots,\tilde{\lambda}^M)U$, where *diag* indicates that the various $\tilde{\lambda}$ are arranged as the elements of the main diagonal of matrix Φ;
wherein each column of matrix Φ is used as a normalized weight vector indicated by $\Phi = [z_1,\ldots,z_N]$ and said normalized weight vectors are made up of individual normalized weights z, $z_i = [z_{i1},\ldots,z_{iN}]$, where i is an integer ranging from 1 to N;
developing unnormalized weight vector $w_i = [w_{i1},\ldots,w_{iN}]$, with each of said weights therein being $\sqrt{\tilde{\lambda}^i} z_{ij}$, where j is an integer ranging from 1 to N.

22. A method for determining weights for use in transmitting signals in communications system having a transmitter with N transmit antennas transmitting over a forward channel to a receiver having L re-

ceiver antennas and a reverse channel for communicating from said receiver to said transmitter, in which there may exist correlation in the signals received by two or more of said L receive antennas, the method comprising the steps of:

determining the number of independent signals M that can be transmitted from said N transmit antennas to said L receive antennas through a process of determining weights for substreams derived from data to be transmitted via said N antennas as part of forming said signals, wherein said weights are determined by solving a matrix equation $H^{\dagger}(K^{N})H = U^{\dagger}\Lambda^{2}U$ where:

H is a channel response matrix,
$H^{\dagger}$ is a conjugate transpose of said channel response matrix H,
K^{N} is the interference covariance matrix,
U is a unitary matrix, each column of which is an eigenvector of $H^{\dagger}(K^{N})H$,
Λ is a diagonal matrix defined as $\Lambda = diag(\lambda^{1},\ldots,\lambda^{M})$, where $\lambda^{1},\ldots,\lambda^{M}$ are each eigenvalues of $H^{\dagger}(K^{N})H$, M being the maximum number of nonzero eigenvalues, which corresponds to the number of said independent signals, and
$U^{\dagger}$ is the conjugate transpose of matrix U;
waterfilling said eigenvalues λ by solving the simultaneous equations $\tilde{\lambda}^{k} = \left(\nu - \frac{1}{(\lambda^{k})^{2}}\right)^{+}$ and $\sum_{k} \tilde{\lambda}^{k} = P$, for ν, where:
k is an integer index that ranges from 1 to M,
P is the transmitted power,
+ is an operator that returns zero (0) when its argument is negative, and returns the argument itself when it is positive, and
each $\tilde{\lambda}$ is an intermediate variable representative of a power for each weight vector;

defining matrix Φ as $\Phi = U^{\dagger} diag(\tilde{\lambda}^{1},\ldots,\tilde{\lambda}^{M})U$, where *diag* indicates that the various $\tilde{\lambda}$ are arranged as the elements of the main diagonal of matrix Φ;
wherein each column of matrix Φ is used as a normalized weight vector indicated by $\Phi = [z_{1},\ldots,z_{N}]$ and said normalized weight vectors are made up of individual normalized weights z, $z_{i} = [z_{i1},\ldots,z_{iN}]$, where i is an integer ranging from 1 to N;
developing unnormalized weight vector $w_{i} = [w_{i1},\ldots, w_{iN}]$, with each of said weights therein being $\sqrt{\tilde{\lambda}^{i}}z_{ij}$, where j is an integer ranging from 1 to N.

FIG. 1

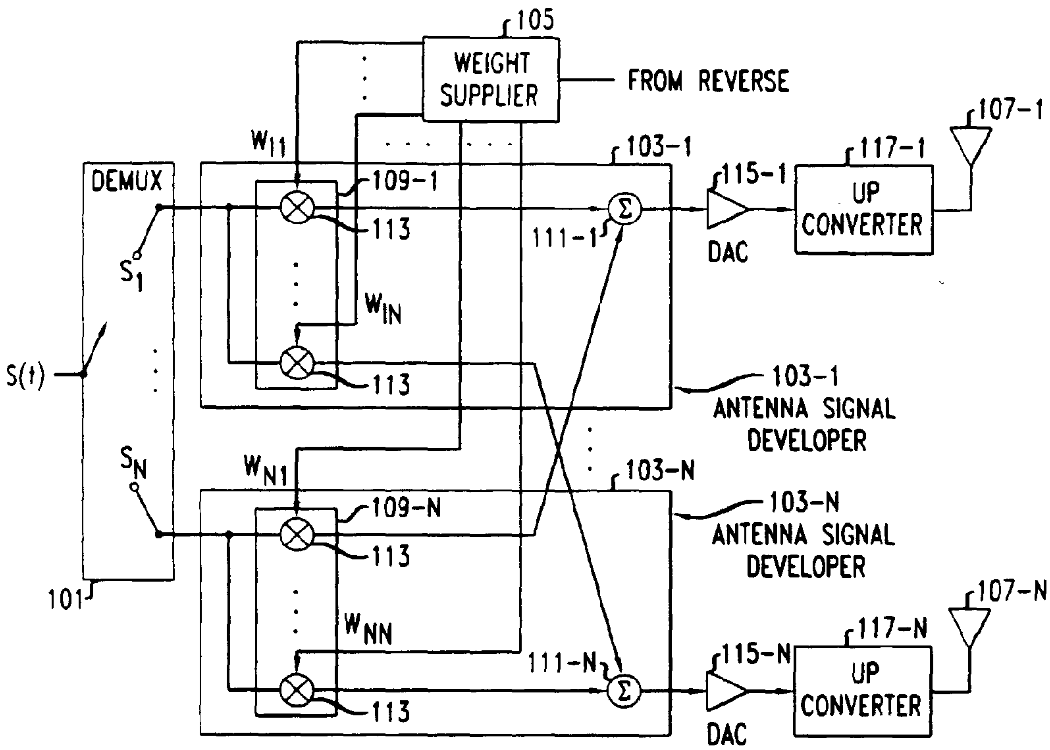

105
WEIGHT SUPPLIER
FROM REVERSE
107-1
W11
103-1
117-1
DEMUX
109-1
115-1
UP CONVERTER
113
111-1
DAC
S1
WIN
113
S(t)
103-1
ANTENNA SIGNAL DEVELOPER
SN
WN1
103-N
109-N
103-N
ANTENNA SIGNAL DEVELOPER
113
101
107-N
WNN
117-N
111-N
115-N
UP CONVERTER
113
DAC

FIG. 2

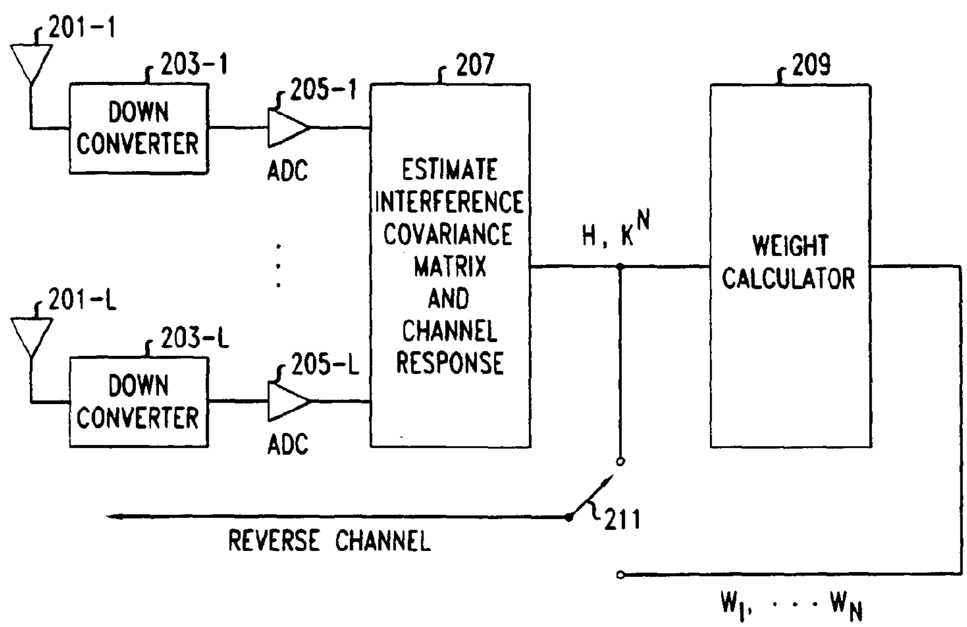

201-1
203-1
207
209
DOWN CONVERTER
205-1
ADC
ESTIMATE INTERFERENCE COVARIANCE MATRIX AND CHANNEL RESPONSE
H, K^N
WEIGHT CALCULATOR
201-L
203-L
205-L
DOWN CONVERTER
ADC
211
REVERSE CHANNEL
W1, · · · WN

FIG. 3

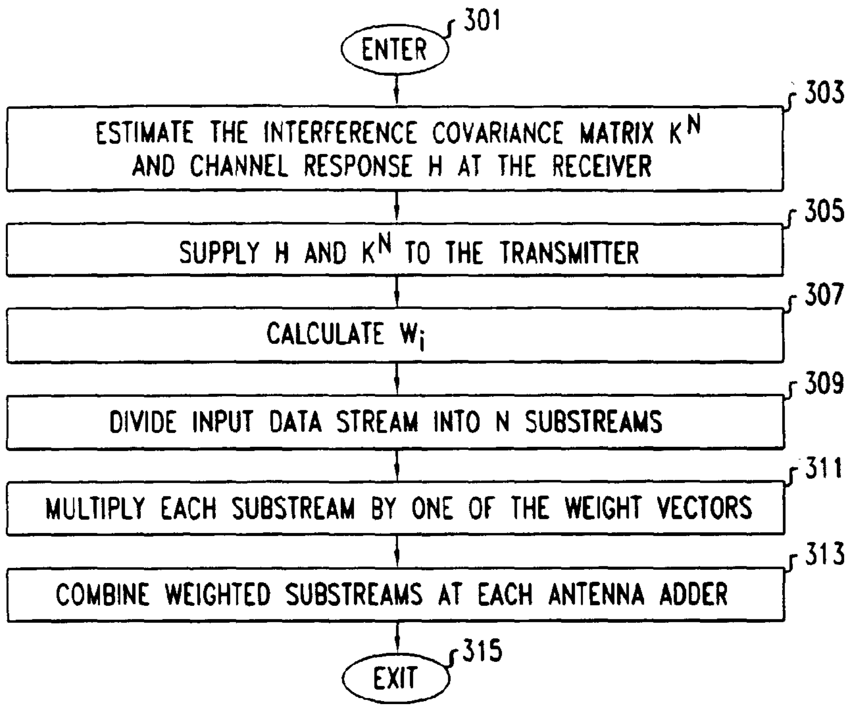
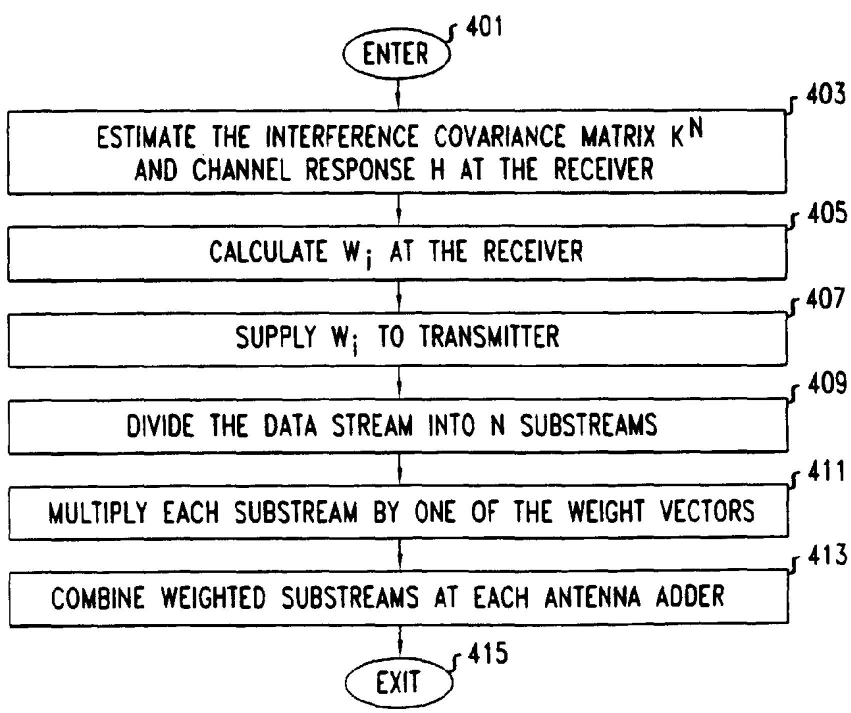

301
ENTER
303
ESTIMATE THE INTERFERENCE COVARIANCE MATRIX K^N AND CHANNEL RESPONSE H AT THE RECEIVER
305
SUPPLY H AND K^N TO THE TRANSMITTER
307
CALCULATE W_i
309
DIVIDE INPUT DATA STREAM INTO N SUBSTREAMS
311
MULTIPLY EACH SUBSTREAM BY ONE OF THE WEIGHT VECTORS
313
COMBINE WEIGHTED SUBSTREAMS AT EACH ANTENNA ADDER
315
EXIT

FIG. 4


401
ENTER
403
ESTIMATE THE INTERFERENCE COVARIANCE MATRIX K^N AND CHANNEL RESPONSE H AT THE RECEIVER
405
CALCULATE W_i AT THE RECEIVER
407
SUPPLY W_i TO TRANSMITTER
409
DIVIDE THE DATA STREAM INTO N SUBSTREAMS
411
MULTIPLY EACH SUBSTREAM BY ONE OF THE WEIGHT VECTORS
413
COMBINE WEIGHTED SUBSTREAMS AT EACH ANTENNA ADDER
415
EXIT

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(71) Applicant: MATSUSHITA ELECTRIC INDUSTRIAL CO., LTD.
Kadoma-shi, Osaka 571-8501 (JP)

(72) Inventor: Goto, Masahiko
Yokosuka-shi, Kanagawa 239-0847 (JP)

(74) Representative: Grünecker, Kinkeldey,
Stockmair & Schwanhäusser Anwaltssozietät
Maximilianstrasse 58
80538 München (DE)

(54) Transmit power control method and apparatus in a radio communication system

(57) When the SIR of the received signal is larger than the SIR upper limit, threshold determining section 105 outputs an instruction for decreasing the transmit power to control bit generating section 106. Further, when the SIR is smaller than the SIR lower limit, threshold determining section 105 outputs an instruction for increasing the transmit power to control bit generating section 106. When the SIR of the received signal is equal to or smaller than the SIR upper limit and equal to or larger than the SIR lower limit, threshold determining section 105 outputs the determined result to control bit generating section 106. When the SIR of the received signal is equal to or smaller than the SIR upper limit and equal to or larger than the SIR lower limit, control bit generating section 106 refers to a last instruction on the transmit power control stored in storage section 107, and outputs an instruction opposite to the last instruction with respect to an increase or decrease in the transmit power to storage section 107 and multiplexing section 108.

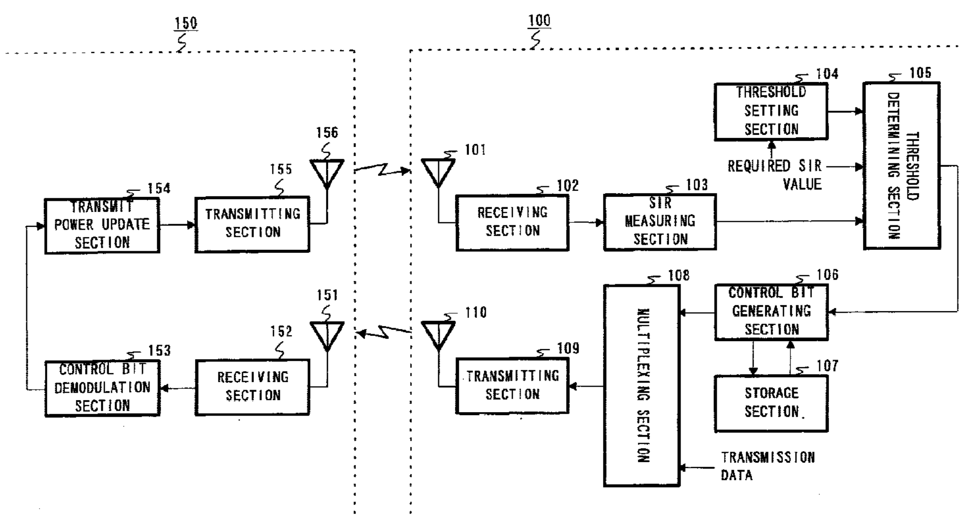

FIG. 3

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Description

BACKGROUND OF THE INVENTION

Field of the Invention

[0001] The present invention relates to a communication apparatus and transmit power control method, and more particularly, to a communication apparatus and transmit power control method used in a radio communication using a CDMA system as a radio access system.

Description of the Related Art

[0002] Recently, in a CDMA radio communication system, a case sometimes occurs that a plurality of control channels and user channels exists together in the same carrier frequency. Therefore, the existence of a channel on which signals are transmitted with the transmit power more than a required level causes the interference with other users, and thereby degrades the reception performance in the other users. Accordingly, it is necessary to suppress the transmit power to an extent enabling a predetermined received quality to be maintained.

[0003] A conventional transmit power control is herein explained. FIG.1 is a diagram illustrating a communication aspect of a base station apparatus and communication terminal apparatus.

[0004] Communication terminal apparatus 10 measures SIR (Signal to Interference Ratio) of a signal transmitted from base station apparatus 20. Then, communication terminal apparatus 10 compares the measured SIR with a required SIR value, and instructs base station apparatus 20 to increase or decrease the transmit power.

[0005] According to the instruction transmitted from communication terminal apparatus 10, base station apparatus 20 changes the transmit power of a signal to be transmitted to communication terminal apparatus 10.

[0006] However, as described in a variable rate CDMA transmit power control method disclosed in Japanese Laid Open Patent Publication HEI11-17646, in a conventional configuration, a transmit power control error sometimes occurs due to a time delay between the time the transmit power control information is determined and the time the transmit power is actually updated according to the transmit power control information.

[0007] An example of the above transmit power control will be described below. FIG.2 is a view showing an example of the conventional transmit power control.

[0008] In FIG.2, the abscissa is indicative of time, the ordinate is indicative of SIR, and instructions for communication terminal apparatus 10 to transmit to base station apparatus 20 are indicated at a lower portion in the figure. In FIG.2, the instruction "UP" is an instruction for increasing the transmit power of base station apparatus 20, while the instruction "DW" is another instruction for decreasing the transmit power of base station apparatus 20.

[0009] S1 is indicative of SIR of a signal received by communication terminal apparatus 10 at time t1. Since S1 has a value lower than a required SIR value a1, communication terminal apparatus 10 transmits the instruction "UP" to base station apparatus 20.

[0010] S2 is indicative of SIR of a signal received by communication terminal apparatus 10 at time t2. Since S2 has a value lower than the required SIR value a1, communication terminal apparatus 10 transmits the instruction "UP" to base station apparatus 20. Base station apparatus 20 receives at time t2 the instruction "UP" transmitted from communication terminal apparatus 10 at time t1, and increases the transmit power to transmit a signal to communication terminal apparatus 10.

[0011] S3 is indicative of SIR of a signal received by communication terminal apparatus 10 at time t3. Since S3 has a value lower than the required SIR value a1, communication terminal apparatus 10 transmits the instruction "UP" to base station apparatus 20. Base station apparatus 20 receives at time t3 the instruction "UP" transmitted from communication terminal apparatus 10 at time t2, and increases the transmit power to transmit a signal to communication terminal apparatus 10.

[0012] S4 is indicative of SIR of a signal received by communication terminal apparatus 10 at time t4. Since S4 has a value lower than the required SIR value a1, communication terminal apparatus 10 transmits the instruction "UP" to base station apparatus 20. Base station apparatus 20 receives at time t4 the instruction "UP" transmitted from communication terminal apparatus 10 at time t3, and increases the transmit power to transmit a signal to communication terminal apparatus 10.

[0013] S5 is indicative of SIR of a signal received by communication terminal apparatus 10 at time t5. Since S5 has a value higher than the required SIR value a1, communication terminal apparatus 10 transmits the instruction "DW" to base station apparatus 20. Base station apparatus 20 receives at time t5 the instruction "UP" transmitted from communication terminal apparatus 10 at time t4, and increases the transmit power to transmit a signal to communication terminal apparatus 10.

[0014] S6 is indicative of SIR of a signal received by communication terminal apparatus 10 at time t6. Since S6 has a value higher than the required SIR value a1, communication terminal apparatus 10 transmits the instruction "DW" to base station apparatus 20. Base station apparatus 20 receives at time t6 the instruction "DW" transmitted from communication terminal apparatus 10 at time t5, and decreases the transmit power to transmit a signal to communication terminal apparatus 10.

[0015] Thus, in communications between communication terminal apparatus 10 and base station apparatus 20, the transmit power control error occurs due to a time delay between the time the transmit power control

information is determined and the time the transmit power is actually updated according to the transmit power control information.

[0016] In the case where the transmit power control information is bit information for instructing an increase or decrease amount in the transmit power, even when the propagation environment of radio signals is static, a time difference arises until the transmit power control information is reflected in a transmit power level, due to a propagation delay of a control signal, and therefore the transmit power level changes more than the increase or decrease amount in the transmit power, and consequently varies widely from the required level. As a result, the SIR value of a signal received by the communication terminal apparatus also varies widely from the required SIR value. In particular, as the delay in the control time is increased, the variation amount in the SIR of the received signal is increased.

[0017] Further, the transmit power control error also occurs under the environment of fading. As a result of the above transmit power variation, the received quality in the communication terminal apparatus deteriorates when the transmit power is controlled below the required power level, while the received qualities in other users may deteriorate when the transmit power is controlled above the required power level.

SUMMARY OF THE INVENTION

[0018] It is an object of the present invention to provide a communication apparatus and transmit power control method that reduce a variation more than an increase or decrease amount in the transmit power in the vicinity of a required level of the transmit power.

[0019] In the closed-loop transmit power control, there occurs a delay between transmitting an instruction on the transmit power control obtained from a received signal quality to a communication partner and reflecting the instruction in the transmit power of the communication partner. Therefore, the present invention achieves the above object by providing a required received quality, which is a criterion for the transmit power control, with a range allowing a transmit power level varying during such a delay to judge, and thereby performing the transmit power control.

[0020] Specifically, the present invention achieves the above object by when the received quality of a received signal is in the range of the required received quality, referring to previous instructions on the transmit power control, instructing the transmit power so that the instructions on the transmit power do not lean in either direction, and thereby reducing the variation in the transmit power due to the propagation delay of a control signal.

[0021] More specifically, the present invention achieves the above object by in the closed-loop transmit power control method, setting a required range from the required level of the received quality, instructing a change opposite to a previously instructed change with respect to the increase or decrease in the transmit power control when the received quality is in the required range, and thereby reducing the variation in the transmit power due to the propagation delay of the control signal.

BRIEF DESCRIPTION OF THE DRAWINGS

[0022] The above and other objects and features of the invention will appear more fully hereinafter from a consideration of the following description taken in connection with the accompanying drawing wherein one example is illustrated by way of example, in which;

FIG.1 is a diagram illustrating a communication aspect of a base station apparatus and a communication terminal apparatus;
FIG.2 is a view showing an example of conventional transmit power control;
FIG.3 is a block diagram illustrating an example of configurations of a communication terminal apparatus and a base station apparatus according to a first embodiment of the present invention;
FIG.4 is a view showing an example of transmit power control between the communication terminal apparatus and base station apparatus according to the first embodiment of the present invention;
FIG.5 is a block diagram illustrating another example of configurations of the communication terminal apparatus and the base station apparatus according to the first embodiment of the present invention;
FIG.6 is a block diagram illustrating another example of configurations of the communication terminal apparatus and the base station apparatus according to the first embodiment of the present invention; and
FIG.7 is a block diagram illustrating an example of configurations of a communication terminal apparatus and a base station apparatus according to a second embodiment of the present invention.

DETAILED DESCRIPTION OF THE PREFERRED EMBODIMENTS

[0023] Embodiments of the present invention will be described below with reference to accompanying drawings.

(First embodiment)

[0024] The first embodiment of present invention explains a case that in the closed-loop transmit power control, the transmit power control is performed by considering a delay occurring during a period while an instruction on transmit power control obtained from a received signal quality is transmitted to a communication partner and then the instruction is reflected in the transmit power of the communication partner, and providing a required

received quality, which is a criterion for the transmit power control, with a range allowing a transmit power level varying during such a delay to judge.

[0025] FIG.3 is a block diagram illustrating an example of configurations of a communication terminal apparatus and a base station apparatus according to the first embodiment of the present invention.

[0026] Communication terminal apparatus 100 is mainly comprised of antenna 101, receiving section 102, SIR measuring section 103, threshold setting section 104, threshold determining section 105, control bit generating section 106, storage section 107, multiplexing section 108, transmitting section 109 and antenna 110.

[0027] Further, base station apparatus 150 is mainly comprised of antenna 151, receiving section 152, control bit demodulation section 153, transmit power update section 154, transmitting section 155, and antenna 156.

[0028] Generally, a delay occurs during a period while communication terminal apparatus 100 transmits an instruction on the transmit power control to base station apparatus 150 and then base station apparatus 150 transmits a signal with the transmit power with the instruction reflected therein to communication terminal apparatus 100, and therefore the received quality changes by an amount corresponding to the time delay in this loop.

[0029] Hence, the required received quality is provided with a range in advance, and when a received quality is in the range, communication terminal apparatus 100 refers to a previous instruction on the transmit power control, and with respect to an instruction on an increase or decrease in the transmit power, transmits an instruction opposite to the previous instruction to base station apparatus 150.

[0030] Antenna 101 receives a radio signal transmitted from base station apparatus 150 to output to receiving section 102. Receiving section 102 converts the received radio signal into a signal with a baseband frequency to demodulate, and outputs the demodulated signal to SIR measuring section 103.

[0031] SIR measuring section 103 measures SIR (Signal to Interference Ratio) of the demodulated signal to output to threshold determining section 105.

[0032] Threshold setting section 104 calculates an SIR upper limit obtained by adding a predetermined amount to the required SIR value and an SIR lower limit obtained by subtracting a predetermined value from the required SIR value to output to threshold determining section 105.

[0033] When the SIR of the received signal is larger than the SIR upper limit, threshold determining section 105 outputs an instruction for decreasing the transmit power to control bit generating section 106. Further, when the SIR is smaller than the SIR lower limit, threshold determining section 105 outputs an instruction for increasing the transmit power to control bit generating section 106.

[0034] Furthermore, when the SIR of the received signal is equal to or smaller than the SIR upper limit and equal to or larger than the SIR lower limit, threshold determining section 105 outputs the determined result to control bit generating section 106.

[0035] According to the determined result in threshold determining section 105, control bit generating section 106 generates a TPC (Transmit Power Control) bit for instructing the transmit power control for a communication partner to output to multiplexing section 108.

[0036] When the SIR of the received signal is equal to or smaller than the SIR upper limit and equal to or larger than the SIR lower limit, control bit generating section 106 refers to a last instruction on the transmit power control stored in storage section 107, and outputs an instruction opposite to the last instruction with respect to an increase or decrease in the transmit power to storage section 107 and multiplexing section 108.

[0037] For example, when an instruction for "decreasing the transmit power by 1dB" is last output, control bit generating section 106 next outputs an instruction for "increasing the transmit power by 1dB" to storage section 107 and multiplexing section 108. Meanwhile, when an instruction for "increasing the transmit power by 1dB" is last output, control bit generating section 106 next outputs an instruction for "decreasing the transmit power by 1dB" to storage section 107 and multiplexing section 108.

[0038] Further, when the SIR value is larger than the SIR upper limit, control bit generating section 106 generates a TPC bit indicative of an instruction for decreasing the transmit power to output to storage section 107 and multiplexing section 108. Meanwhile, when the SIR value is smaller than the SIR lower limit, control bit generating section 106 generates a TPC bit indicative of an instruction for increasing the transmit power to output to storage section 107 and multiplexing section 108.

[0039] Storage section 107 stores the instructions on the transmit power control generated in control bit generating section 106.

[0040] Multiplexing section 108 multiplexes transmission data and the TPC bit to output to transmitting section 109. Transmitting section 109 modulates the multiplexed transmission data and TPC bit, converts the resultant signal into a radio signal, and transmits the radio signal through antenna 110.

[0041] Receiving section 152 receives through antenna 151 the radio signal transmitted from communication terminal apparatus 100, and converts the radio signal into a baseband signal to output to control bit demodulation section 153. Control bit demodulation section 153 extracts a transmit power control bit from the received signal to output to transmit power update section 154. Transmit power update section 154 changes the transmit power according to the information contained in the transmit power control bit.

[0042] Thus, communication terminal apparatus 100 measures the SIR of a signal transmitted from base sta-

tion apparatus 150 to judge the transmit power control, and transmits an instruction on the transmit power control to base station apparatus 150. Then, according to the instruction on the transmit power control transmitted from communication terminal apparatus 100, base station apparatus 150 changes the transmit power.

[0043] The operation on the transmit power control will be explained next. FIG.4 is a diagram illustrating an example of the transmit power control between the communication terminal apparatus and base station apparatus according to the first embodiment of the present invention.

[0044] In FIG.4, the abscissa is indicative of time, the ordinate is indicative of SIR, and instructions for communication terminal apparatus 100 to transmit to base station apparatus 150 are indicated at a lower portion in the figure. In FIG.4, the instruction "UP" is an instruction for increasing the transmit power of base station apparatus 150, while the instruction "DW" is another instruction for decreasing the transmit power of base station apparatus 150.

[0045] Communication terminal apparatus 100 sets a required SIR value a1, an SIR lower limit a2, and an SIR upper limit a3.

[0046] Changes in SIR of received signals and instructions on the transmit power control will be explained below with respect to time t1 to t6 in this order.

[0047] S1 is indicative of SIR of a signal received by communication terminal apparatus 100 at time t1. Since S1 has a value lower than the SIR lower limit a2, communication terminal apparatus 100 transmits the instruction "UP" at time t1.

[0048] S2 is indicative of SIR of a signal received by communication terminal apparatus 100 at time t2. Since S2 has a value lower than the SIR lower limit a2, communication terminal apparatus 100 transmits the instruction "UP" at time t2. Base station apparatus 150 receives at time t2 the instruction "UP" transmitted from communication terminal apparatus 100 at time t1, and increases the transmit power to transmit a signal to communication terminal apparatus 100.

[0049] S3 is indicative of SIR of a signal received by communication terminal apparatus 100 at time t3. Since S3 has a value lower than the SIR lower limit a2, communication terminal apparatus 100 transmits the instruction "UP" at time t3. Base station apparatus 150 receives at time t3 the instruction "UP" transmitted from communication terminal apparatus at time t2, and increases the transmit power to transmit a signal to communication terminal apparatus 100.

[0050] S4 is indicative of SIR of a signal received by communication terminal apparatus 100 at time t4. Since S4 has a value equal to or higher than the SIR lower limit a2 and equal to or lower than the SIR upper limit a3, communication terminal apparatus 100 refers to the last transmitted instruction, in other words, the instruction "UP" transmitted at time t3, and transmits the instruction "DW" opposite to the instruction "UP" at time t4.

[0051] Base station apparatus 150 receives at time t4 the instruction "UP" transmitted from communication terminal apparatus at time t3, and increases the transmit power to transmit a signal to communication terminal apparatus 100.

[0052] S5 is indicative of SIR of a signal received by communication terminal apparatus 100 at time t5. Since S5 has a value equal to or higher than the SIR lower limit a2 and equal to or lower than the upper limit a3, communication terminal apparatus 100 refers to the last transmitted instruction, in other words, the instruction "DW" transmitted at time t4, and transmits the instruction "UP" opposite to the instruction "DW" at time t5.

[0053] Base station apparatus 150 receives at time t5 the instruction "DW" transmitted from communication terminal apparatus 100 at time t4, and decreases the transmit power to transmit a signal to communication terminal apparatus 100.

[0054] S6 is indicative of SIR of a signal received by communication terminal apparatus 100 at time t6. Since S6 has a value equal to or higher than the SIR lower limit a2 and equal to or lower than the upper limit a3, communication terminal apparatus 100 refers to the last transmitted instruction, in other words, the instruction "UP" transmitted at time t5, and transmits the instruction "DW" opposite to the instruction "UP" to base station apparatus 150.

[0055] Base station apparatus 150 receives at time t6 the instruction "UP" transmitted from communication terminal apparatus 100 at time t5, and increases the transmit power to transmit a signal to communication terminal apparatus 100.

[0056] Thus, according to the communication terminal apparatus and base station apparatus of the first embodiment of the present invention, in the closed-loop transmit power control, it is possible to reduce the variation more than an increase or decrease amount in the transmit power in the vicinity of the required level of the transmit power, by setting a required range from the required level of the received quality, and when the received quality is in the required range, instructing a change opposite to a previously instructed change with respect to the increase or decrease in the transmit power control.

[0057] In addition, it is preferable that the required range of the received quality is more than or equal to a sum of a transmit power change amount multiplied by a delay time and a transmit power change amount at one time. That is, the required range of the received quality is only required to be more than a amount changing in the transmit power during a delay between transmitting a transmit power control instruction and reflecting the instruction.

[0058] Further, the communication terminal apparatus of this embodiment is capable of changing the threshold into an arbitrary value at an arbitrary timing. FIG.5 is a block diagram illustrating another example of

configurations of the communication terminal apparatus and the base station apparatus in the first embodiment of the present invention. In addition, sections common to those in FIG.3 are assigned the same reference numerals as in FIG.3, and the detailed explanation is omitted.

[0059] Communication terminal apparatus 200 in FIG.5 is different from the communication terminal apparatus in FIG.3 in a point that the apparatus 200 is provided with threshold setting section 201 to change a threshold into an arbitrary value at an arbitrary timing.

[0060] In FIG.5, threshold setting section 201 outputs an instruction for changing a threshold into an arbitrary value at an arbitrary timing to threshold determining section 105. For example, when the signal delay is large and the delay between transmitting a transmit power instruction and reflecting the instruction is increased, threshold setting section 201 outputs an instruction for increasing the SIR upper limit and another instruction for decreasing the SIR lower limit to threshold determining section 105.

[0061] Further, when the signal delay is small and the delay between transmitting a transmit power instruction and reflecting the instruction is decreased, threshold setting section 201 outputs an instruction for decreasing the SIR upper limit and another instruction for increasing the SIR lower limit to threshold determining section 105.

[0062] According to the instructions output from threshold setting section 201, threshold determining section 105 changes the SIR upper limit and SIR lower limit to compare with the SIR of a received signal.

[0063] Thus, communication terminal apparatus 200 of this embodiment changes the SIR upper limit and SIR lower limit corresponding to a delay in the propagation path, and thereby makes a range from the SIR upper limit to the SIR lower limit more than a change amount in the transmit power control due to the propagation delay. As a result, when the SIR value is in the vicinity of the required SIR value, it is possible to prevent the transmit power from being changed more than a change in the transmit power during a period from the time the received quality is compared to the time the instruction on the transmit power is reflected.

[0064] Further, the communication terminal apparatus of this embodiment is capable of changing the SIR upper limit and SIR lower limit according to a change of the required SIR value. FIG.6 is a block diagram illustrating another example of configurations of the communication terminal apparatus and the base station apparatus in the first embodiment of the present invention. In addition, sections common to those in FIG.3 are assigned the same reference numerals as in FIG.3, and the detailed explanation is omitted.

[0065] Communication terminal apparatus 300 in FIG.6 is different from the communication terminal apparatus in FIG.3 in a point that the apparatus 300 is provided with threshold setting section 301 to change the SIR upper limit and SIR lower limit according to a change of the required SIR value.

[0066] In FIG.6, when the required SIR value is changed, threshold setting section 301 outputs an instruction for changing the SIR upper limit and the SIR lower limit by the same increase or decrease amount as in changing the required SIR value. According to the instruction output from threshold setting section 301, threshold determining section 105 changes the SIR upper limit and the SIR lower limit to compare with SIR of a received signal.

[0067] Further, during a period or a plurality of periods of measuring SIR immediately after changing the required SIR value, the threshold determining section 301 instructs to decrease the transmit power when the measured SIR value is more than or equal to the required SIR value, while instructing to increase the transmit power when the measured SIR value is less than the required SIR value.

[0068] Thus, communication terminal apparatus 300 of this embodiment changes the SIR upper limit and the SIR lower limit by the same increase or decrease amount as in changing the required SIR value, whereby the apparatus 300 is capable of reducing a variation more than an increase or decrease amount in the transmit power in the vicinity of the required level of the transmit power even when the required SIR value is changed.

(Second embodiment)

[0069] FIG.7 is a block diagram illustrating an example of configurations of a communication terminal apparatus and a base station apparatus according to the second embodiment of the present invention.

[0070] In FIG.7, communication terminal apparatus 400 is mainly comprised of antenna 401, receiving section 402, SIR measuring section 403, threshold setting section 404, threshold determining section 405, control bit generating section 406, storage section 407, multiplexing section 408, control bit demodulation section 409, transmit power update section 410, transmitting section 411 and antenna 412.

[0071] Further, base station apparatus 450 is mainly comprised of antenna 451, receiving section 452, SIR measuring section 453, threshold setting section 454, threshold determining section 455, control bit generating section 456, storage section 457, multiplexing section 458, control bit demodulation section 459, transmit power update section 460, transmitting section 461, and antenna 462.

[0072] Antenna 401 receives a radio signal transmitted from base station apparatus 450 to output to receiving section 402. Receiving section 402 converts the received radio signal into a signal with a baseband frequency to demodulate, and outputs the demodulated signal to SIR measuring section 403 and control bit demodulation section 409.

[0073] SIR measuring section 403 measures SIR of the demodulated signal to output to threshold determin-

ing section 405. Threshold setting section 404 calculates an SIR upper limit obtained by adding a predetermined amount to the required SIR value and an SIR lower limit obtained by subtracting a predetermined value from the required SIR value to output to threshold determining section 405.

[0074] When the SIR of the received signal is larger than the SIR upper limit, threshold determining section 405 outputs an instruction for decreasing the transmit power to control bit generating section 406. Further, when the SIR is smaller than the SIR lower limit, threshold determining section 405 outputs an instruction for increasing the transmit power to control bit generating section 406.

[0075] Furthermore, when the SIR of the received signal is equal to or smaller than the SIR upper limit and equal to or larger than the SIR lower limit, threshold determining section 405 outputs the determined result to control bit generating section 406. According to the determined result in threshold determining section 405, control bit generating section 406 generates a TPC bit for instructing the transmit power control for a communication partner to output to multiplexing section 408.

[0076] When the SIR of the received signal is equal to or smaller than the SIR upper limit and equal to or larger than the SIR lower limit, control bit generating section 406 refers to a last instruction on the transmit power control stored in storage section 407, and outputs an instruction opposite to the last instruction with respect to an increase or decrease in the transmit power to storage section 407 and multiplexing section 408. The specific operation of control bit generating section 406 is the same as in control bit generating section 106 of the first embodiment.

[0077] Further, when the SIR of the received signal is larger than the SIR upper limit, control bit generating section 406 generates a TPC bit indicative of an instruction for decreasing the transmit power to output to storage section 407 and multiplexing section 408. Meanwhile, when the SIR of the received signal is smaller than the SIR lower limit, control bit generating section 406 generates a TPC bit indicative of an instruction for increasing the transmit power to output to storage section 407 and multiplexing section 408.

[0078] Storage section 407 stores the instructions on the transmit power control generated in control bit generating section 406. Multiplexing section 408 multiplexes transmission data and the TPC bit to output to transmitting section 411.

[0079] Control bit demodulation section 409 extracts a transmit power control bit from the received signal to output to transmit power update section 410. Transmit power update section 410 changes the transmit power according to the information contained in the transmit power control bit. Transmitting section 411 modulates the multiplexed transmission data and TPC bit, converts the resultant signal to a radio signal, and transmits the radio signal through antenna 412.

[0080] Antenna 451 receives the radio signal transmitted from communication terminal apparatus 400 to output to receiving section 452. Receiving section 452 converts the received radio signal into a signal with a baseband frequency to demodulate, and outputs the demodulated signal to SIR measuring section 453 and control bit demodulation section 459.

[0081] SIR measuring section 453 measures SIR of the demodulated signal to output to threshold determining section 455. Threshold setting section 454 calculates an SIR upper limit obtained by adding a predetermined amount to the required SIR value and an SIR lower limit obtained by subtracting a predetermined value from the required SIR value to output to threshold determining section 455.

[0082] When the SIR of the received signal is larger than the SIR lower limit, threshold determining section 455 outputs an instruction for decreasing the transmit power to control bit generating section 456. Further, when the SIR is smaller than the SIR upper limit, threshold determining section 455 outputs an instruction for increasing the transmit power to control bit generating section 456.

[0083] Furthermore, when the SIR of the received signal is equal to or smaller than the SIR upper limit and equal to or larger than the SIR lower limit, threshold determining section 455 outputs the determined result to control bit generating section 456. According to the determined result in threshold determining section 455, control bit generating section 456 generates a TPC bit for instructing the transmit power control for a communication partner to output to multiplexing section 458.

[0084] When the SIR of the received signal is equal to or smaller than the SIR upper limit and equal to or larger than the SIR lower limit, control bit generating section 456 refers to a last instruction on the transmit power control stored in storage section 457, and outputs an instruction opposite to the last instruction with respect to an increase or decrease in the transmit power to storage section 457 and multiplexing section 458. The specific operation of control bit generating section 456 is the same as in control bit generating section 106 of the first embodiment.

[0085] Further, when the SIR is larger than the SIR upper limit, control bit generating section 456 generates a TPC bit indicative of an instruction for decreasing the transmit power to output to storage section 457 and multiplexing section 458. Meanwhile, when the SIR is smaller than the SIR lower limit, control bit generating section 456 generates a TPC bit indicative of an instruction for increasing the transmit power to output to storage section 457 and multiplexing section 458.

[0086] Storage section 457 stores the instructions on the transmit power control generated in control bit generating section 456. Multiplexing section 458 multiplexes transmission data and the TPC bit to output to transmitting section 461.

[0087] Control bit demodulation section 459 extracts

a transmit power control bit from the received signal to output to transmit power update section 460. Transmit power update section 460 changes the transmit power according to the information contained in the transmit power control bit. Transmitting section 461 modulates the multiplexed transmission data and TPC bit, converts the resultant signal to a radio signal, and transmits the radio signal through antenna 462.

[0088] Thus, communication terminal apparatus 400 measures the SIR of a signal transmitted from base station apparatus 450 to judge the transmit power control, and transmits an instruction on the transmit power control to base station apparatus 450. Then, according to the instruction on the transmit power control transmitted from communication terminal apparatus 400, base station apparatus 450 changes the transmit power.

[0089] Further, base station apparatus 450 measures the SIR of a signal transmitted from communication terminal apparatus 400 to judge the transmit power control, and transmits an instruction on the transmit power control to communication terminal apparatus 400. Then, according to the instruction on the transmit power control transmitted from base station apparatus 450, communication terminal apparatus 400 changes the transmit power.

[0090] Thus, according to communication terminal apparatus 400 and base station apparatus 450 of the second embodiment of the present invention, in the closed-loop transmit power control, it is possible to reduce the variation more than an increase or decrease amount in the transmit power in the vicinity of the required level of the transmit power, by setting a required range from the required level of the received quality, and when the received quality is in the required range, instructing a change opposite to a previously instructed change with respect to the increase or decrease in the transmit power control.

[0091] In an apparatus of the communication partner, the above processing is performed in the same way, whereby the transmit power control information to be transmitted and a transmit power level of a signal to be transmitted are determined.

[0092] In addition, while this embodiment describes about a communication terminal apparatus and/or base station apparatus, the present invention is not limited to those, and any communication apparatuses are applicable that control transmit power of a communication partner from a received quality of a signal transmitted from the communication partner.

[0093] Further, while the embodiments of the present invention explain a case that the SIR of a received signal is compared with the required SIR, the present invention is not limited to the above case, and any values are applicable that are indicative of a received quality.

[0094] The transmit power control apparatus in the CDMA radio communication as described above is capable of reducing a variation in the measured SIR value in the vicinity of the required SIR value due to a delay in control time. As a result, the apparatus is capable of properly controlling the transmit power level to an extent enabling a predetermined received quality to be maintained and further of decreasing the interference in other users.

[0095] As this invention may be embodied in several forms without departing from the spirit of essential characteristics thereof, the present embodiment is therefore illustrative and not restrictive, since the scope of the invention is defined by the appended claims rather than by the description preceding them, and all changes that fall within meets and bounds of the claims, or equivalence of such meets and bounds are therefore intended to embraced by the claims.

[0096] The present invention is not limited to the above described embodiments, and various variations and modifications may be possible without departing from the scope of the present invention.

[0097] This application is based on the Japanese Patent Application No.2000-057195 filed on March 2, 2000, entire content of which is expressly incorporated by reference herein.

Claims

1. A radio communication apparatus comprising:

    measuring means (103) for measuring a received quality of a signal transmitted from a communication partner;
    instructing means (105,106) for performing an instruction for changing transmit power to said communication partner from a result obtained by comparing the received quality of the signal with a required received quality;
    storage means (107) for storing the instruction for changing transmit power; and
    transmitting means (109) for transmitting the instruction for changing transmit power to said communication partner,

    wherein said instructing means provides a value of the required received quality with a range allowing a change in the transmit power caused by a propagation delay of a transmit power control signal to perform the instruction for changing the transmit power.

2. The radio communication apparatus according to claim 1, wherein said instructing means (105,106) performs the instruction for decreasing the transmit power to said communication partner when the received quality of the signal is higher than an upper limit of a predetermined range, further performs the instruction for increasing the transmit power to said communication partner when the received quality of the signal is lower than a lower limit of the prede-

termined range, and refers to a previous instruction for changing the transmit power stored in said storage means (107) to perform the instruction for changing the transmit power so as to maintain the received quality of the signal in the predetermined range when the received quality of the signal is in the predetermined range.

3. The radio communication apparatus according to claim 2, wherein said instructing means (105,106) refers to the previous instruction for changing the transmit power stored in said storage means (107) when the received quality of the signal is in the predetermined range, and when a last instruction for changing transmit power is indicative of increasing the transmit power, generates the instruction for decreasing the transmit power, while when the last instruction for changing transmit power is indicative of decreasing the transmit power, generating the instruction for increasing the transmit power.

4. The radio communication apparatus according to claim 3, further comprising:

setting means (104) for setting the predetermined range with a predetermined variation range from the required received quality.

5. The radio communication apparatus according to claim 1, wherein said instructing means (105,106) changes the range into an arbitrary range at an arbitrary timing.

6. The radio communication apparatus according to claim 1, wherein when the required received quality is changed, said instructing means (105,106) performs the instruction for decreasing the transmit power when the received quality of the signal is higher than the required received quality, while performing the instruction for increasing the transmit power when the received quality of the signal is lower than the required received quality, not depending on whether the received quality of the signal is in the range during one or a plurality of measuring periods immediately after changing the received quality.

7. A mobile station apparatus having a radio communication apparatus, said radio communication apparatus comprising:

measuring means (103) for measuring a received quality of a signal transmitted from a communication partner;
instructing means (105,106) for generating an instruction for changing transmit power to said communication partner from a result obtained by comparing the received quality of the signal with a required received quality;
storage means (107) for storing the instruction for changing transmit power; and
transmitting means (109) for transmitting the instruction for changing transmit power to said communication partner,

wherein said instructing means performs the instruction for decreasing the transmit power to said communication partner when the received quality of the signal is higher than an upper limit of a predetermined range, further performs the instruction for increasing the transmit power to said communication partner when the received quality of the signal is lower than a lower limit of the predetermined range, and refers to a previous instruction for changing the transmit power stored in said storage means to perform the instruction for changing the transmit power so as to maintain the received quality of the signal in the predetermined range when the received quality of the signal is in the predetermined range.

8. A base station apparatus having a radio communication apparatus, said radio communication apparatus comprising:

measuring means (103) for measuring a received quality of a signal transmitted from a communication partner;
instructing means (105,106) for generating an instruction for changing transmit power to said communication partner from a result obtained by comparing the received quality of the signal with a required received quality;
storage means (107) for storing the instruction for changing transmit power; and
transmitting means (109) for transmitting the instruction for changing transmit power to said communication partner,

wherein said instructing means performs the instruction for decreasing the transmit power to said communication partner when the received quality of the signal is higher than an upper limit of a predetermined range, further performs the instruction for increasing the transmit power to said communication partner when the received quality of the signal is lower than a lower limit of the predetermined range, and refers to a previous instruction for changing the transmit power stored in said storage means to perform the instruction for changing the transmit power so as to maintain the received quality of the signal in the predetermined range when the received quality of the signal is in the predetermined range.

9. A transmit power control method, comprising:

measuring a received quality of a signal transmitted from a communication partner;
comparing the received quality of the signal with a required received quality, and based on the compared result, performing an instruction for changing transmit power to said communication partner
storing the instruction for changing transmit power; and
transmitting the instruction for decreasing the transmit power to said communication partner when the received quality of the signal is higher than an upper limit of a predetermined range, further transmitting the instruction for increasing the transmit power to said communication partner when the received quality of the signal is lower than a lower limit of the predetermined range, and referring to a stored previous instruction for changing the transmit power to transmit the instruction for changing the transmit power such that the received quality of the signal is maintained in the predetermined range to said communication partner when the received quality of the signal is in the predetermined range.

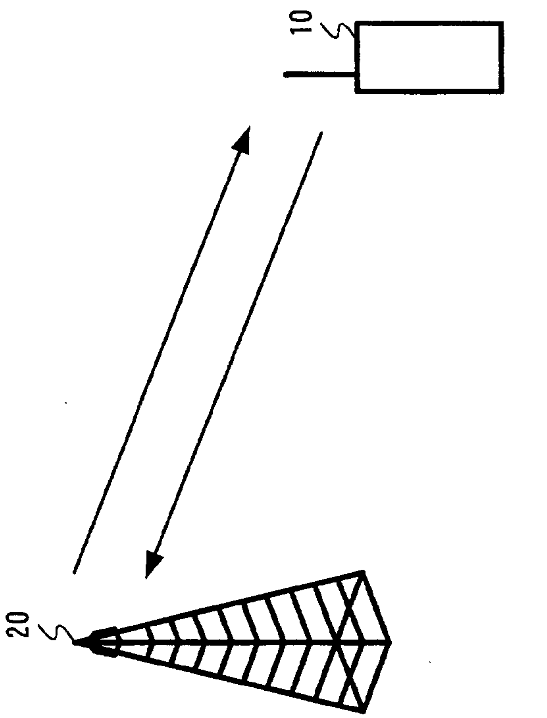
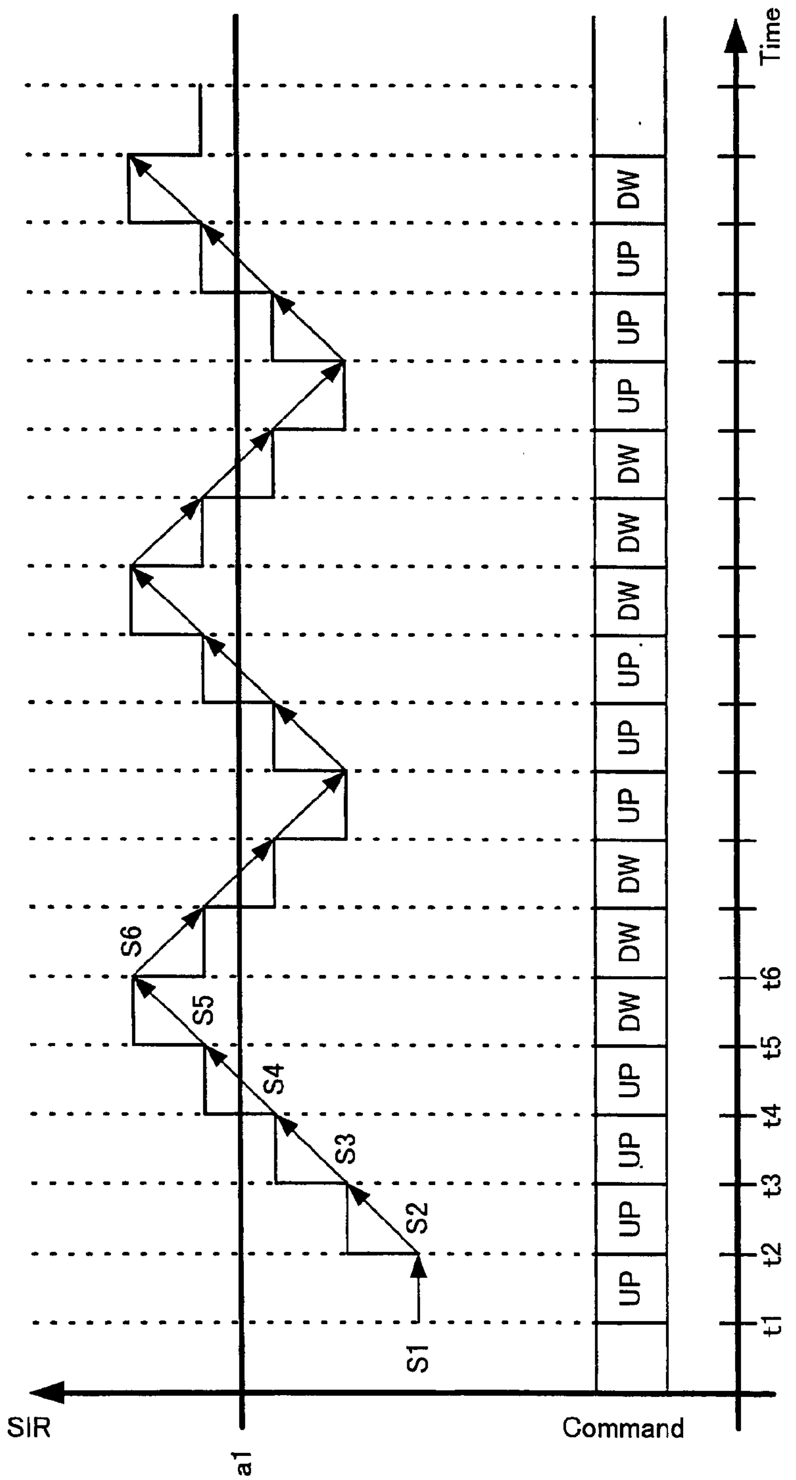


FIG. 1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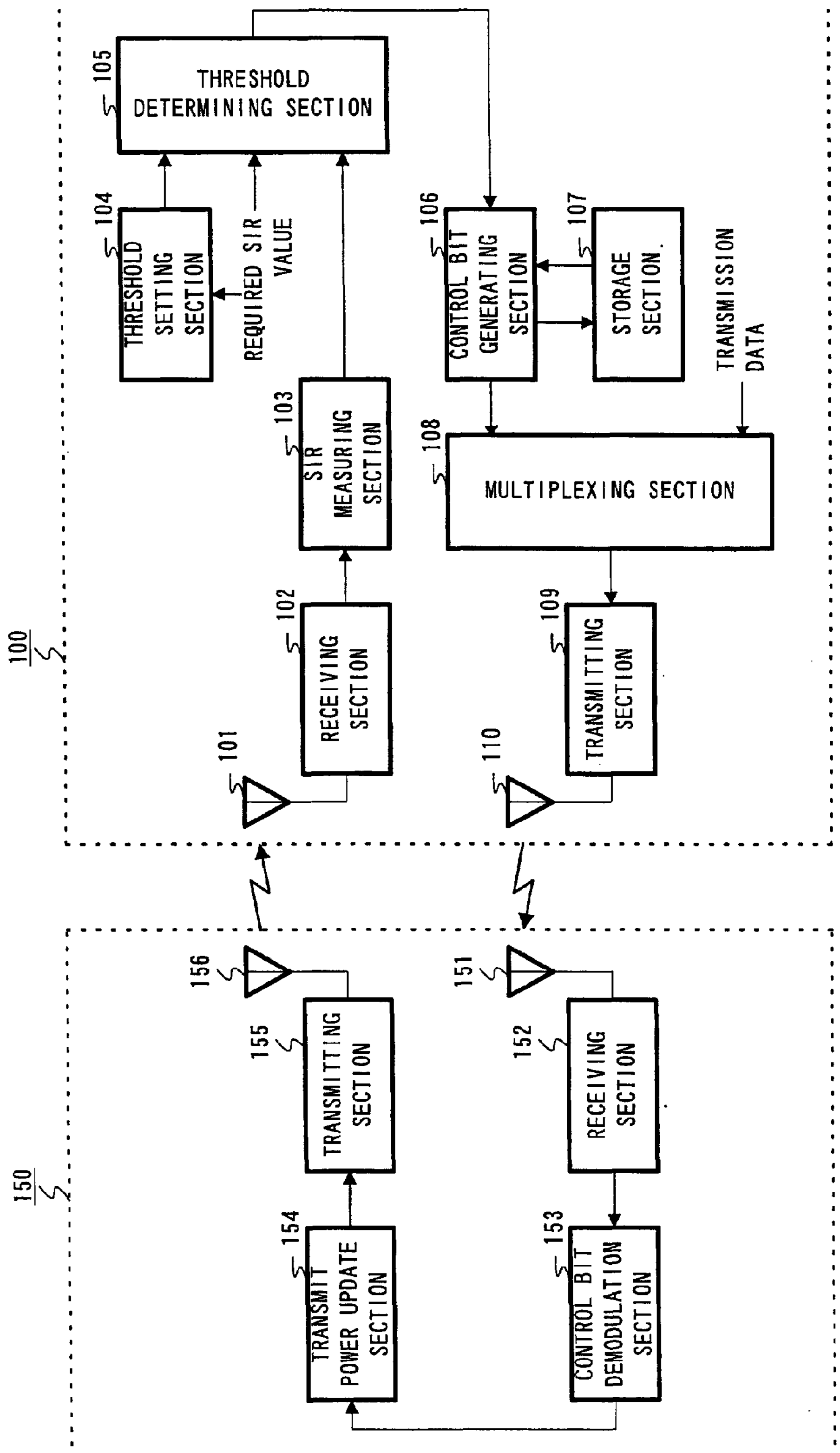
UP
DW

FIG. 2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THRESHOLD SETTING SECTION
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105
CONTROL BIT GENERATING SECTION
106
STORAGE SECTION
107
MULTIPLEXING SECTION
108
TRANSMISSION DATA
TRANSMITTING SECTION
109
110
150
151
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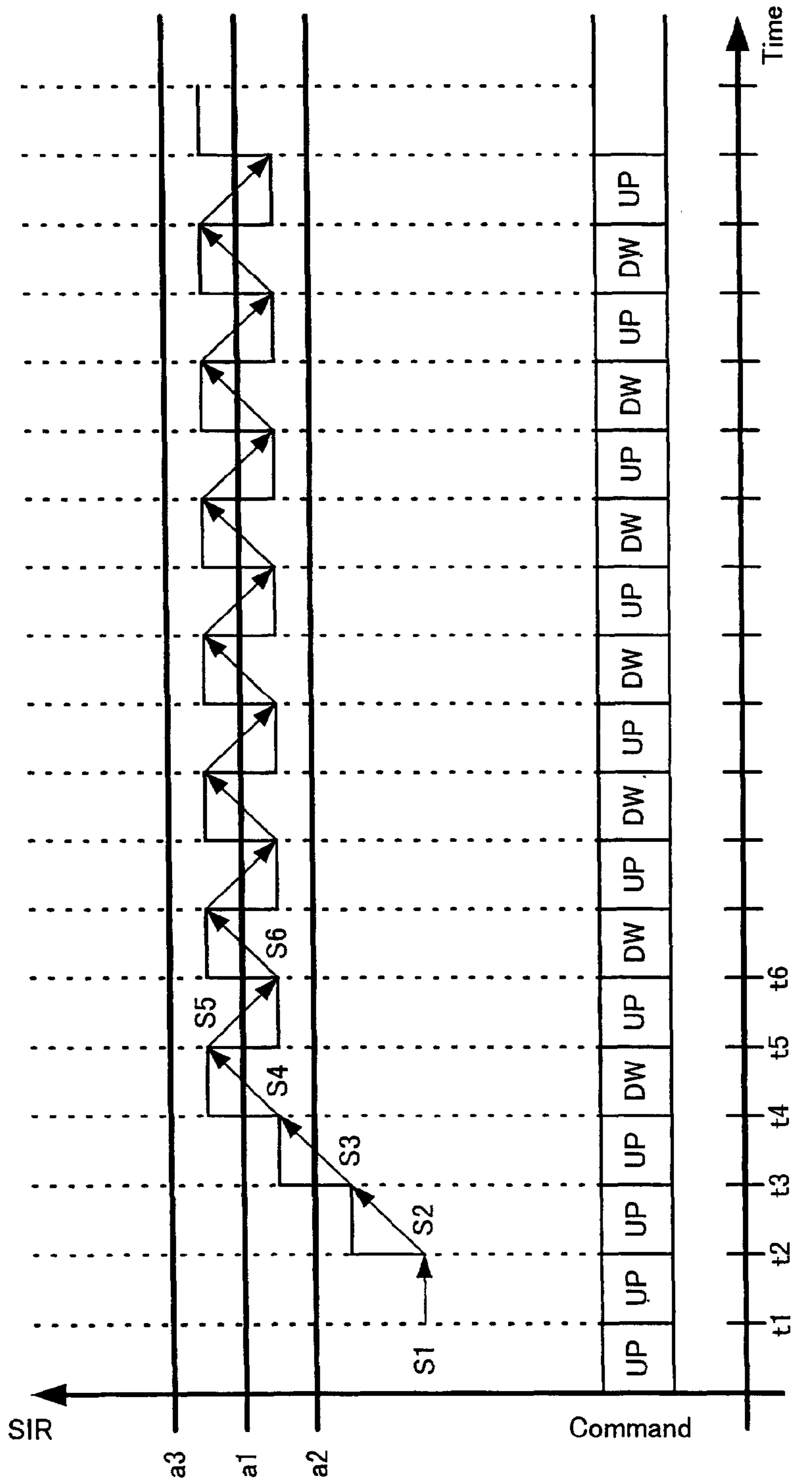
CONTROL BIT DEMODULATION SECTION
153
TRANSMIT POWER UPDATE SECTION
154
TRANSMITTING SECTION
155
156

FIG. 3



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S2
S3
S4
S5
S6
Command
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UP
UP
DW
UP
DW
UP
DW
UP
DW
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t1
t2
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t5
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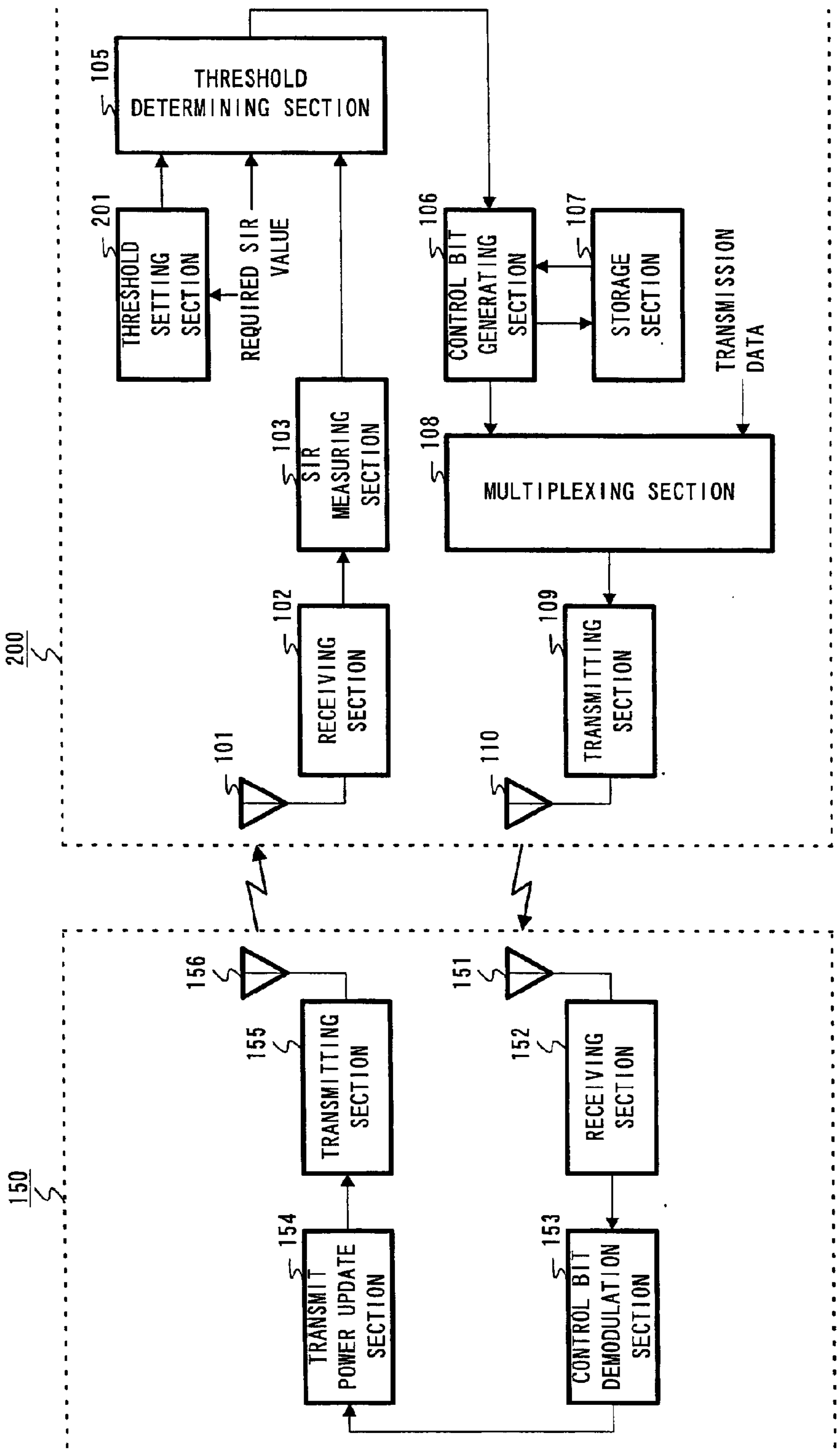
Time

F I G. 4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106
CONTROL BIT GENERATING SECTION
107
STORAGE SECTION
TRANSMISSION DATA
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MULTIPLEXING SECTION
109
TRANSMITTING SECTION
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RECEIVING SECTION
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CONTROL BIT DEMODULATION SECTION

FIG. 5

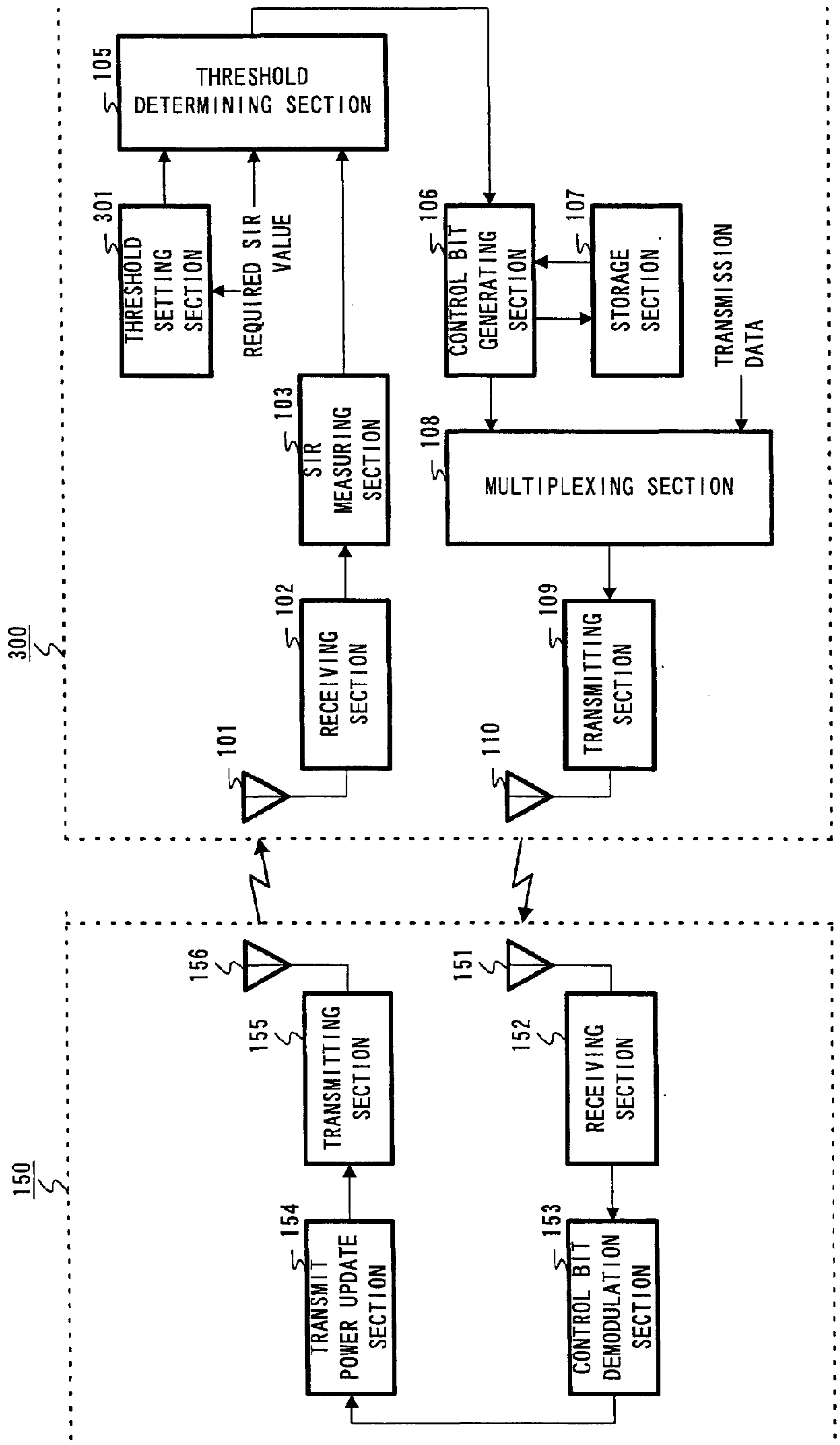
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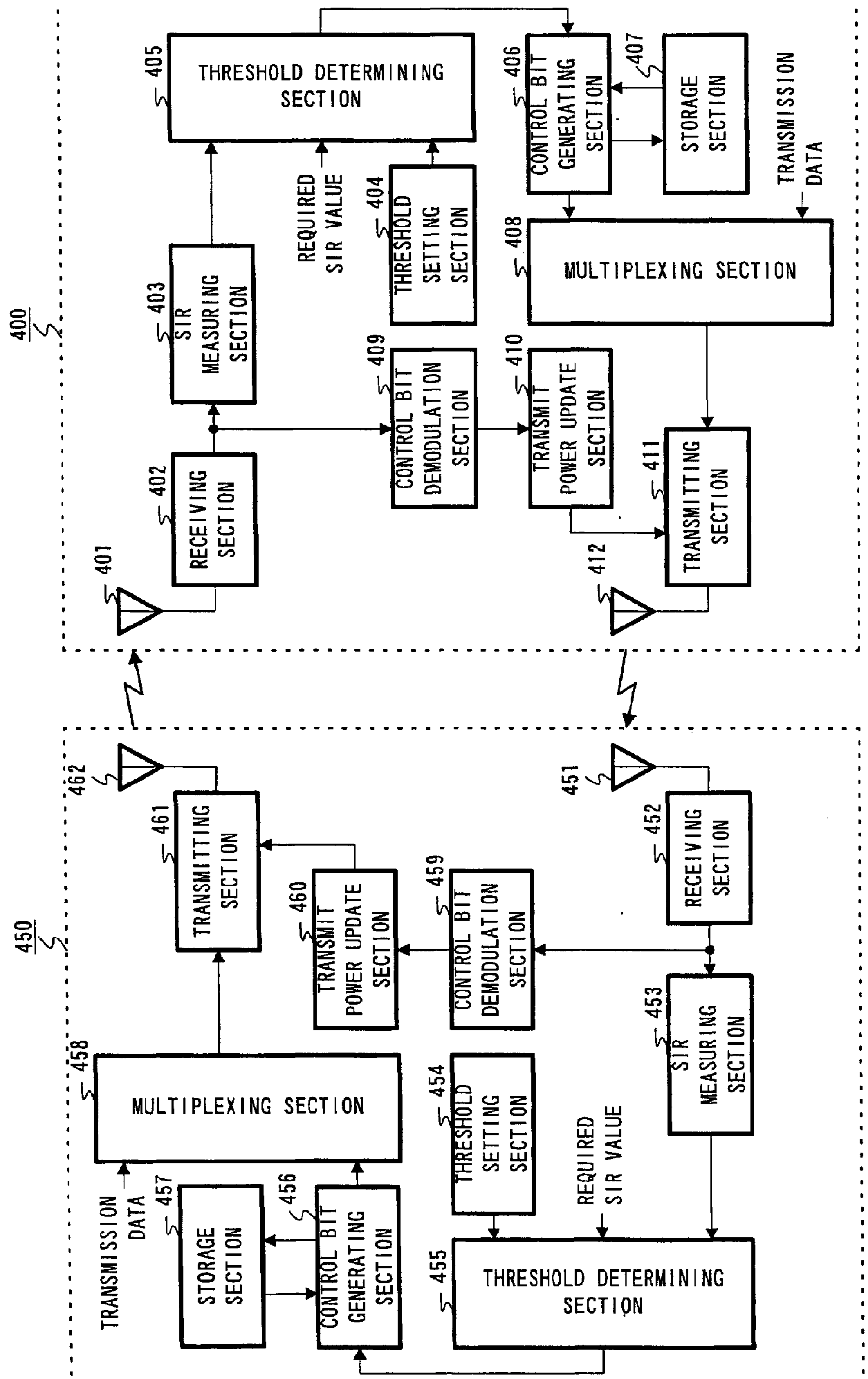
FIG. 6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CONTROL BIT GENERATING SECTION
406
STORAGE SECTION
407
TRANSMISSION DATA
MULTIPLEXING SECTION
408
CONTROL BIT DEMODULATION SECTION
409
TRANSMIT POWER UPDATE SECTION
410
TRANSMITTING SECTION
411
412
450
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RECEIVING SECTION
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SIR MEASURING SECTION
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THRESHOLD SETTING SECTION
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REQUIRED SIR VALUE
THRESHOLD DETERMINING SECTION
455
CONTROL BIT GENERATING SECTION
456
STORAGE SECTION
457
TRANSMISSION DATA
MULTIPLEXING SECTION
458
CONTROL BIT DEMODULATION SECTION
459
TRANSMIT POWER UPDATE SECTION
460
TRANSMITTING SECTION
461
462

FIG. 7

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(71) Applicant: Sony International (Europe) GmbH
10785 Berlin (DE)

(72) Inventors:
- Kornprobst, Stefan,
Telecommunications R&D Europe
70327 Stuttgart (DE)
- Jürgensen, Jens-Uwe,
Telecommunications R&D Europe
70327 Stuttgart (DE)
- Stirling-Gallacher, Richard, Comm. R&D Europe
70327 Stuttgart (DE)

(74) Representative: Körber, Martin, Dipl.-Phys.
Mitscherlich & Partner
Patentanwälte
Sonnenstrasse 33
80331 München (DE)

(54) Event triggered change of access service class in a random access channel

(57) The present invention relates to a communication device (1) for transmitting and receiving data in a communication system and a corresponding communication method. In the communication system, which can be a wireless or a wired communication system, a random access channel is provided, which comprises a plurality of access resources. The access resources are divided in at least two access resource groups, each access resource groups corresponding to a different access service class with a respective access probability. The communication device (1) according to the present invention, comprises selecting means (5) for randomly selecting an access resource from an access resource group corresponding to the current access service class of the communication device (1), transmitting means (3) for transmitting a random access burst in said selected access resource and detecting means (6) for detecting a specific event, whereby the current access service class of the communication device (1) is changed into another access service class when said specific event is detected by the detecting means (6).

The present invention proposes a simple and flexible scheme for the adaptation of the performance of a communication device (1) in the random access procedure in the case of certain events.

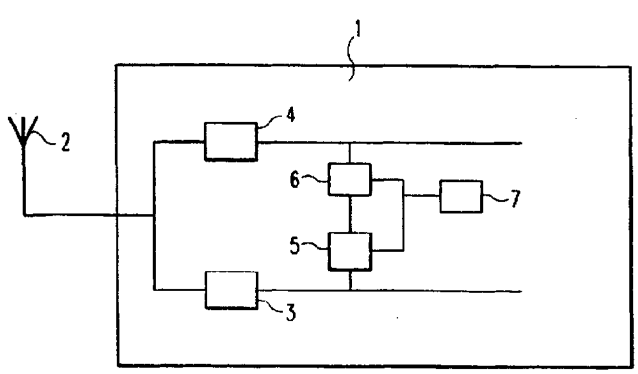

Fig. 1

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Description

[0001] The present invention relates to a communication device for transmitting and receiving data in a communication system and to a communication method for such a communication device. In the communication system, a random access channel is provided, whereby said random access channel provides a plurality of access resources. A random access channel is a communication channel used to build up a connection from one communication device to another communication device of the communication system. The communication system can thereby be a wireless communication system or a wired communication system or a mixture of both.

[0002] The access resources of the random access channel are randomly accessed by a communication device which for example wants to build up a connection or to send a short message. Thereby, the access is contention based which means that several communication devices trying to build up a connection might try to access the same access resource simultaneously. In wireless telecommunication systems, as for example the UMTS-system or a UMTS-like system, where a mobile station wanting to build up a connection and/or transmit data requests an access resource of an random access channel by randomly selecting an access resource and transmitting a preamble part of a random access burst to the base station. Hereby, the preamble part represents the request for the randomly selected access resource. A base station receiving the preamble part grants the requested access resource if it is available and sends a corresponding acknowledgement signal back to the mobile station. In a certain situation, for example, if no appropriate access resource is available on the network side, the base station transmits a negative acknowledgement signal back to the mobile station indicating that the requested access resource is not available. In case that the respective base station grants the request, it transmits a positive acknowledgement signal. The mobile station receiving the positive acknowledgement signal then transmits the message part of the random access burst, which contains the data upon which the building up of a connection or the transmission of data between the mobile station and the base station is based.

[0003] In case that several mobile terminals access the same access resource simultaneously, the base station is not able to receive the access requests and thus does not transmit any acknowledgement signal. The mobile terminals then have to send access requests again.

[0004] Depending on the respective communication system, the access resources may be time slots as in an TDMA (Time Divisional Multiple Access) system, frequencies as in a FDMA (Frequency Division Multiple Access) system, codes as in a CDMA (Code Division Multiple Access) system or mixtures thereof, as in the UMTS system, in which an access resource in the random access channel is defined by a time slot and a signature code.

[0005] The general problem is that the mechanism of gaining access to the access resources of the random access channel is contention based. In order to allow a more efficient access to the random access channel, different proposals have been made. For example, EP 98 118 819.6 which is a document according to article 54 (3) EPC, proposes to divide the access resources of the random access channel into several groups, whereby each group corresponds to a respective priority class. Each priority class thereby represents the transmission priority of the random access data to be transmitted. Thus, the access requests for different kinds of data are transmitted with different priorities, so that certain kinds of data have a higher probability to gain access to an access resource of the random access channel than other kinds of data. This proposed system, however, is very inflexible, since a respective number of access resources is fixedly allocated to a specific kind of data. Further, the only criteria for the allocation of a specific access priority and thus the access probability is the data kind. WO 97/19525 proposes a more flexible system, in which the access of a random access channel in a communication system relies on the use of varying access probabilities for subscribers or messages of varying priority. Thereby, users, for example mobile terminals, are divided into different priority groups, whereby each group has a different access probability for accessing an access resource of the random access channel. Two basic schemes of the distribution and allocation of access probabilities to the users are proposed, namely a proportional priority distribution and a temporal priority distribution. In the proportional priority distribution, each group of users attempts access to all available access resources of the random access channel, but with different access probabilities. Each user group has a different access probability, but every access resource can be accessed by every user. In the temporal priority distribution, the highest priority user group, i. e. the user group having the highest access probability to the access resources, attempts access to all available access resources, whereby the lower priority user groups, i. e. the user groups having a lower access probability, only attempt access to a part of the available access resources. The disadvantage here is that access attempts of the highest priority user group have to content with the access attempts of all other priority groups. Thus, it is possible that an access attempt of a very high priority, as for example, an emergency call, is not successful since it has to content with other, lower priority access attempts.

[0006] The object of the present invention is therefore to provide a communication device for transmitting and receiving data in a communication system, and a communication method for a communication device of a communication system, which provides an improved,

more effective and more flexible way of accessing access resources of a random access channel.

[0007] The above object is solved by a communication device for transmitting and receiving data in a communication system according to claim 1, whereby a random access channel is provided, said random access channel providing a plurality of access resources being divided in at least two access resource groups, each access resource group corresponding to a different access service class with a respective access probability. The communication device according to the present invention comprises selecting means for randomly selecting an access resource from an access resource group corresponding to the current access service class of the communication device, transmitting means for transmitting a random access burst in said selected access resource, and detecting means for detecting a specific event, whereby said current access service class of the communication device is changed into another access service class when said specific event is detected by said detecting means. The above object is further achieved by a communication method according to claim 12.

[0008] The present invention therefore provides a very flexible way of accessing access resources of a random access channel. The access service class of a communication device can be changed very easily upon occurrence of a predetermined event so that a higher or lower access probability to the access resources of the random access channel can be chosen depending on the respective requirements. The occurrence of certain events can automatically leads to a change of the quality and the priority of the following accesses depending on predefined rules. Thus, an easy and flexible adaptation of the performance and quality of the access can be achieved. Particularly, an access service class is not fixedly linked to a certain data type, so that the access service class may be changed even if the same type of data is to be transmitted. Advantageous features of the present invention are claimed in the respective subclaims. For example, it may be advantageous, if the access resources of the access resource group corresponding to the random access class having the highest random access probability are exclusively allocated to this access resource group. In this way it can be assured that very important access requests, as for example for an emergency call, do not have to content with other access requests and get an access resource granted very quickly.

[0009] Further advantageously, the access resources of each access resource group can be exclusively allocated to the respective access resource groups. In this case, the access resources of the random access channel are divided into different groups without any overlap between the different groups. Alternatively, it may be advantageous, if some of the access resources are allocated to two or more of the access resource groups.

[0010] Advantageously, the specific event is the reception of a predetermined number of negative acknowledgement signals from another communication device as after sending random access requests on said random access channel. For example, if a certain number of random access requests had been sent unsuccessfully and a predetermined number of negative acknowledgement signals had been received, the current access service class can be changed into another access service class having a higher access probability. Alternatively, it may be advantageous, if said specific event is a time point. For example, the current access service class can be changed into another access service class at certain time points. For example the access service class is changed at time points at which certain system parameters change regularly and a change of the access service classes is required. In this case the current access service class may for example be changed periodically.

[0011] Further advantageously, the rules according to which the current access service class is changed into another access service class are stored in a memory means. For example, the communication device according to the present invention may be a mobile station of a wireless telecommunication system, whereby the memory means is part of a subscriber identity module or a similar device which can be inserted into the mobile station. The rules according to which the current access service class is changed into another access service class may be received from another communication device, as for example a base station of a mobile telecommunication system. In this way, a flexible adaptation to varying system parameters and conditions can be achieved.

[0012] Further advantageously, the communication system is a wireless UMTS telecommunication system, whereby the access resources of the random access channel are defined by time slots and signature codes.

[0013] Advantageous embodiments of the present invention are explained in greater detail in the following description relating to the enclosed drawings, in which

Fig. 1 shows a schematic diagram of a communication device according to the present invention,

Fig. 2 shows a first example of dividing access resources into groups and

Fig. 3 shows a second example of dividing access resources into groups.

[0014] Fig. 1 shows a schematic block diagram of a communication device 1 according to the present invention. The communication device 1 shown in Fig. 1 is a mobile terminal for a wireless telecommunication system and comprises an antenna 2 connected to a transmitting means 3 and a receiving means 4 for transmitting and receiving communication data to and from a base station or another mobile terminal of the wireless tele-

communication system. The wireless telecommunication system can for example be the UMTS system or a UMTS-like system. The mobile terminal 1 thus comprises all necessary elements for communicating and processing data, such as coders, decoders, modulators, demodulators and the like, although these elements are not shown in Fig. 1 and not explained in the present application. It is to be noted, that the mobile terminal 1 shown in Fig. 1 is only used as an example for the communication device according to the present invention, which may also be an end terminal in a wired communication system, such as a telephone apparatus, a personal computer or the like.

[0015] The communication system, in which the communication device 1 according to the present invention operates and in which the communication method according to the present invention is performed, comprises a random access channel (RACH) providing a plurality of access resources. These resources are used by the communication device 1 to build up a connection. In case of a wireless telecommunication system, a mobile terminal uses the access resources of the random access channel to transmit an access request to a corresponding base station. In the UMTS system or a UMTS-like system, the access request is transmitted in the preamble part of the random access burst as explained above. The access resources are thereby accessed randomly, which leads to a contention based access mechanism. Several access requests coming from different mobile terminals may compete or content for the same access resource at the base station. In this case the base station does not receive any access request transmitted from the mobile terminals and the mobile terminals do not receive any acknowledgement signal and have to send the access requests again after a certain time period. The same scheme applies to communication devices which are connected through a wired network.

[0016] According to the present invention, the plurality of access resources of a random access channel is divided into at least two access resource groups. Each access resource group corresponds to a different access service class with a respective access probability. Each communication device of the communication system is allocated to one of the access resource groups and therefore has a current access service class. Each access service class corresponds to a respective access probability so that each communication device of one access resource groups has the same access probability to access one of the access resources of the respective access resource group. Thus, different access service classes can be defined, each access service class having a different access probability. If, for example, three access services classes are defined, a first access service class may have high priority and a high access probability, a second access service class may have a medium priority and a medium access probability and a third access service class may have a low priority and thus a low access probability. In a standard setting, for example, each type of data for a given communication device may be allocated to one of the access resource groups. Each time a specific data type is to be transmitted in the random access channel, a corresponding access probability is valid. According to the present invention, the corresponding access service class is changed upon the occurrence of a specific event.

[0017] The communication device 1 shown in Fig. 1 comprises a selecting means 5 connected to the transmitting means 3. The selecting means 5 randomly selects an access resource from an access resource group corresponding to the current access service class of the communication device 1. Thus, if the communication device 1 or its user wants to build up a connection, an access resource is randomly selected by the selecting means 5 and an access request is transmitted via the transmitting means 5 and the antenna 2 to a corresponding base station. In case that the request is granted, the base station transmits a positive acknowledgement signal which is received by means of the antenna 2 and the receiving means 4 of the communication device 1. The positive acknowledgement signal received by the receiving means 4 is further processed in the communication device 1 and the actual data for building up a connection and/or transmitting data are transmitted thereafter in the granted access resource. These mechanisms are well known and do not need to be described in further detail. An important feature, however, is that the access request already carries the information on the randomly selected access resource so that upon grant of the requested access resource the corresponding connection data are automatically transmitted in that access resource.

[0018] The selecting means 5 always selects the access resource from an access resource group corresponding to the current access service class of the communication device 1. According to the present invention, this current access service class can be changed upon detecting a specific event in a detecting means 6 which is connected to the receiving means 4 and the selecting means 5. In this way, the access probability can be changed to a higher or lower access probability depending on the detected specific event and corresponding rules which are stored in a memory means 7 connected to the detecting means 6 and the selecting means 5. The rules stored in the memory means 7 thereby define to which access service class the current access service class of the communication device 1 is changed upon the detection of a specific event. The specific event is for example the reception of a predetermined number of negative acknowledgement signals after sending an access request. In this case, the current access service class could be changed into another access service class having a higher priority and a higher probability that an access is granted. Alternatively, the specific event could be the immediately successful transmission

of the predetermined number of access grants and a corresponding reception of positive acknowledgement signals. In this case, the current access service class could be changed into a lower priority class having a lower access probability. Other possibilities is to specify the specific event as a time point so that, for example, specific time points could be defined, at which it is statistically known that users of the communication system change their behaviour in respect to the transmission of access requests. Further, the current access service class may be changed periodically. It is to be noted that any kind of a specific event could be defined, upon the detection of which the current access service class is changed. Further, a combination of different kinds of events can- be defined. The change of the access service class is hereby independent from the specific kind of data to be transmitted.

[0019] The rules, according to which the current access service class is changed into another access service class upon detection of a specific event in the detecting means 6 are stored in memory means 7. Thereby, the memory means 7 can be a fixed part of the communication device 1. Alternatively, the memory means 7 can be part of a device which can be inserted into the communication device 1. In this case, the memory means 7 may for example be part of a device which has an inherent association with the user and which is usually inserted into the communication device, as for example a unified subscriber identity module (USIM) card in case of the UMTS system. Further, the rules could be received from another communication device, such as a base station and then stored in the memory means 7. Also, a combination of the above-mentioned possibilities could be realised, for example rules could be stored on a subscriber identity module as well as in a fixed memory means of the communication device 1. In this case, when a subscriber identity module is inserted and connected to the communication device 1, the rule stored in the fixed memory means could be overwritten or overruled by the rules stored in the memory means of the subscriber identity module. Further, new rules received from another communication device could be used to overwrite or overrule the current rule stored on a subscriber identity module and/or a fixed memory means of the communication device 1. In this way, it is possible to flexibly adapt the rules for changing the communication device 1 from a current access service class into another access service class to varying system parameters.

[0020] Figures 2 and 3 show a first and a second example, respectively, of access resources being divided in two ore more access resource groups according to the present invention. Generally, the access resources can be defined by a frequency, a time slot or a code or any combination thereof depending on the multiple access scheme used in the corresponding wireless or wired communication system. Fig. 2 shows an example, in which the access resources of the random access channel are time slots. One repetition cycle of the random access time window comprises for example eight time slots 0, 1,...7. In the example shown in Fig. 2, the available eight access resources or time slots are divided into three access service groups, each corresponding to a different access service class with respective access probability. The first group comprises the five time slots 0,.3,4, 6, 7 indicated by the diagonal pattern, the second access resource group comprises two time slots 2, 5 indicated by the vertical pattern and the third access recourse group comprises one time slot 1, indicated by the horizontal pattern. Communication devices being currently allocated to the first access service class corresponding to the first access resource group therefore have a higher probability to gain access, since there is a higher number and a larger capacity of access resources are available. This is of course only true when approximately an equal number of communication devices is allocated to each access service class. A communication device being currently allocated to the third access service class corresponding to the third access resource group has only a single time slot in each random access time window to gain access and thus has a low probability leading to a poorer performance. Upon detection of a specific event, a communication device belonging to the third access service class may be switched into the second access service class or the first access service class depending on the change rules.

[0021] Fig. 3 shows a second example of access resources being divided into access resource groups according to the present invention. In the shown example, an access resource is defined by a time slot (or time offset) and a preamble signature (or signature code). In a UMTS system, for example, a random access channel comprises up to 15 time slots and 16 preamble signatures within two radio frames or two random access time windows, so that up to 240 access resources are available in total. Fig. 3 shows a corresponding example with eight time slots 0, 1, ... 7 and 16 preamble signatures 0, 1,..... 15. The access resources are divided into three access resource groups. The access resources indicated by a cross from the first access resource group are defined by the time slots 0, 1, 2, 3 and the preamble signatures 0, 1, 2, 3, 4, 5, 6, 7 and correspond to a first access service class. The access resources indicated by a point form of the second access resource group are defined by the time slots 4, 5, 6, 7 and the preamble signatures 0, 1, 2, 3, 4, 5, 6, 7 and correspond to a second access service class. The access resources indicated by a blank square form of the third access research group, are defined by the time slots 0, 1, 2, 3, 4, 5, 6, 7 and the preamble signatures 8, 9, 10, 11, 12, 13, 14, 15 and correspond to a third access service class. In both examples shown in Fig. 2 and 3, the access resources are properly divided into several groups without any overlap between the groups. However, an overlap between the groups is possible, which means that one ore more access resources may belong to two or more

access resource groups.

[0022] The communication device and the communication method according to the present invention provide a simple and effective way of flexibly adapting the performance of a communication device within the random access procedure upon the occurrence of certain events. The proposed scheme deals with changing access service classes and is particularly advantageous for very high priority kind of data, as for example, emergency calls, which can be allocated to a high priority access service class with an exclusive access resource group in which no competition with other non-emergency calls is occurring.

Claims

1. Communication device (1) for transmitting and receiving data in a communication system, in which a random access channel is provided, said random access channel providing a plurality of access resources being divided in at least two access resource groups, each access resource group corresponding to a different access service class with a respective access probability, comprising

   selecting means (5) for randomly selecting an access resource from an access resource group corresponding to the current access service class of the communication device,
   transmitting means (3) for transmitting a random access burst in said selected access resource, and
   detecting means (6) for detecting a specific event, whereby said current access service class of the communication device is changed into another access service class when said specific event is detected by said detecting means (6).

2. Communication device (1) according to claim 1,
   characterized in,
   that the access resources of the access resource group corresponding to the random access class having the highest random access probability are exclusively allocated to this access resource group.

3. Communication device (1) according to claim 1 or 2,
   characterized in,
   that the access resources of each access resource group are exclusively allocated to their respective access resource group.

4. Communication device (1) according to claim 1 or 2,
   characterized in,
   that some access resources are allocated to two or more access resource groups.

5. Communication device (1) according to one of the claims 1 to 4,
   characterized in,
   that rules according to which said current access service class is changed into another access service class are stored in a memory means (7).

6. Communication device (1) according to claim 5,
   characterized in
   being a mobile station of a wireless telecommunication system, whereby said memory means (7) is part of a subscriber identity module.

7. Communication device (1) according to one of the claims 1 to 6,
   characterized in,
   that rules according to which said current access service class is changed into another access service class are received from another communication device.

8. Communication device (1) according to one of the claims 1 to 7,
   characterized in,
   that said specific event is the reception of a predetermined number of negative acknowledgment signals from another communication device after sending random access requests on said random access channel.

9. Communication device (1) according to one of the claims 1 to 7,
   characterized in,
   that said specific event is a time point.

10. Communication device (10) according to one of the claims 1 to 9,
    characterized in,
    that said current access service class is changed periodically.

11. Communication device (1) according to one of the claims 1 to 10,
    characterized in,
    that said communication system is a wireless UMTS telecommunication system, whereby said access resources of said random access channel are defined by time slots and signature codes.

12. Communication method for a communication device (1) of a communication system, in which a random access channel is provided, said random access channel providing a plurality of access resources being divided in at least two access resource groups, each access resource group corresponding to a different access service class with a respective access probability, comprising the steps of

randomly selecting an access resource from an access resource group corresponding to the current access service class of the communication device,
transmitting a random access burst in said selected access resource, and
detecting a specific event, whereby said current access service class of the communication device is changed into another access service class when said specific event is detected.

13. Communication method according to claim 12, **characterized in,**
that the random access resources of the access resource group corresponding to the random access class having the highest random access probability are exclusively allocated to this access resource group.

14. Communication device according to claim 12 or 13, **characterized in,**
that the random access resources of each access resource group are exclusively allocated to their respective access resource group.

15. Communication method according to claim 12 or 13, **characterized in,**
that some random access resources are allocated to two or more access resource groups.

16. Communication method according to one of the claims 12 to 15, **characterized in,**
that rules according to which said current access service class is changed into another access service class are stored in and read from a memory means (7).

17. Communication method according to one of the claims 12 or 16, **characterized in,**
that rules according to which said current access service class is changed into another access service class are transmitted from another communication device of the communication system.

18. Communication method according to one of the claims 12 to 17, **characterized in,**
that said specific event is the reception of a predetermined number of negative acknowledgment signals from another communication device after sending random access requests on said random access channel.

19. Communication method according to one of the claims 12 to 17, **characterized in,**
that said specific event is a time point.

20. Communication method according to claim 19, **characterized in,**
that said current access service class is changed periodically.

21. Communication device according to one of the claims 12 to 20, **characterized in,**
that said communication system is a wireless UMTS telecommunication system, whereby said access resources of said random access channel are defined by time slots and signature codes.

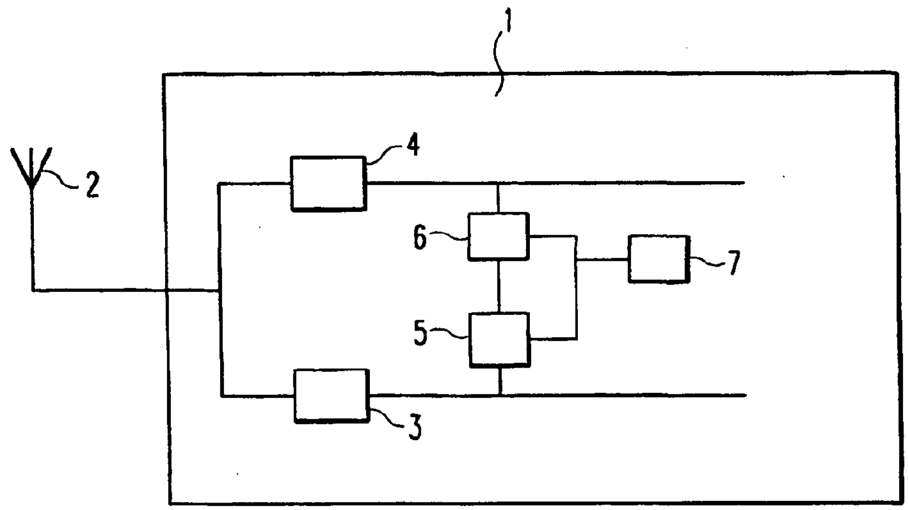

Fig. 1

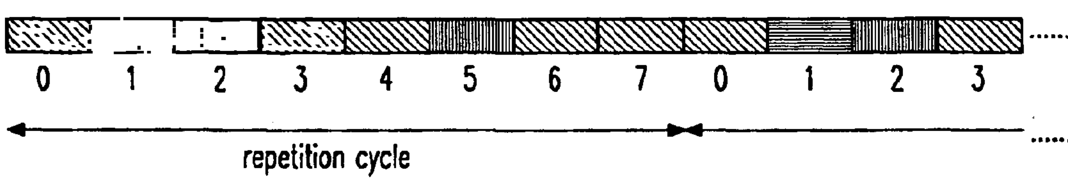

Fig. 2

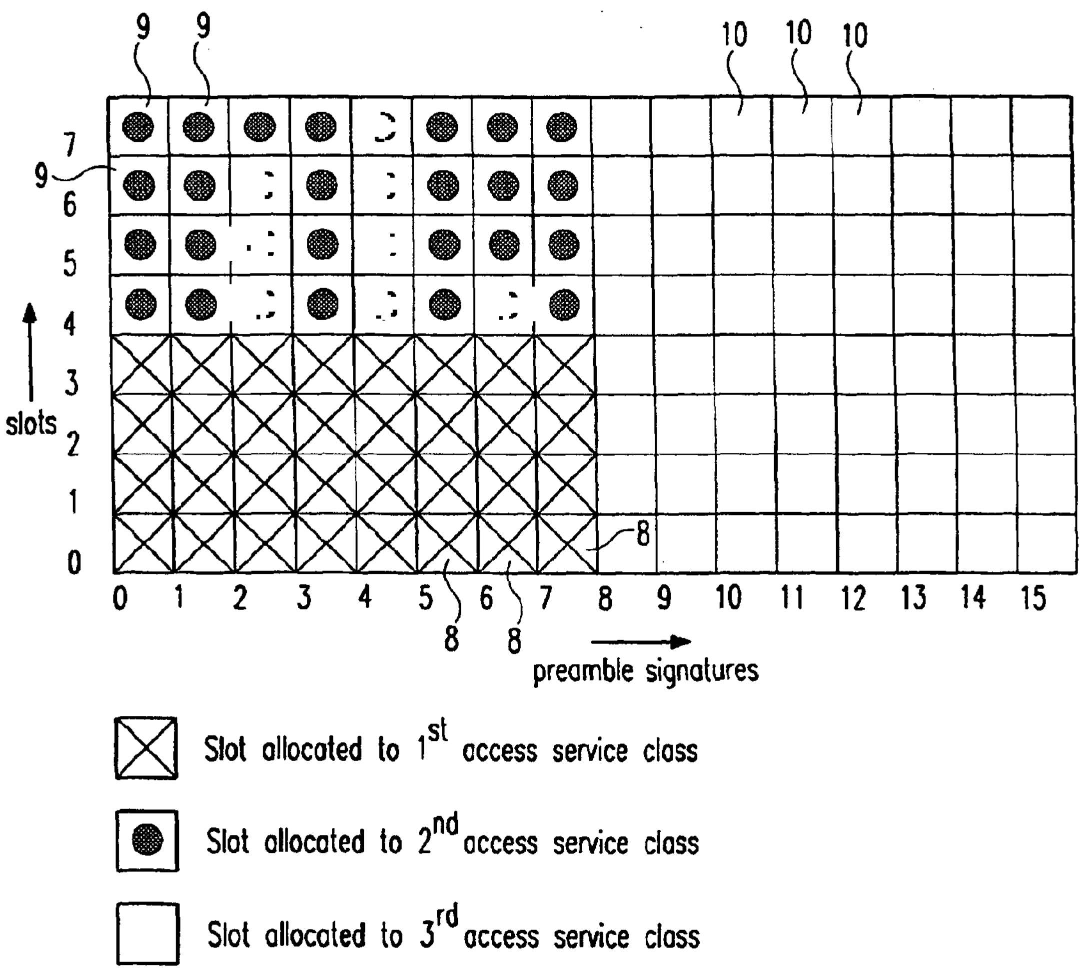
9
9
10
10
10
7
9
6
5
4
3
slots
2
1
0
8
0 1 2 3 4 5 6 7 8 9 10 11 12 13 14 15
8
8
preamble signatures
Slot allocated to 1st access service class
Slot allocated to 2nd access service class
Slot allocated to 3rd access service class

Fig. 3

European Patent Office

EUROPEAN SEARCH REPORT

Application Number
EP 00 10 7329

DOCUMENTS CONSIDERED TO BE RELEVANT

Category	Citation of document with indication, where appropriate, of relevant passages	Relevant to claim	CLASSIFICATION OF THE APPLICATION (Int.Cl.7)
D,A	WO 97 19525 A (MOTOROLA INC) 29 May 1997 (1997-05-29) * page 4, line 2 - line 20 * * page 5, line 25 - page 6, line 19 * * page 11, line 27 - page 13, line 4 * * figure 6 *	1-4, 12-15	H04Q7/38
A	WO 99 44379 A (ERICSSON TELEFON AB L M) 2 September 1999 (1999-09-02) * page 4, line 2 - page 5, line 3 * * page 16, line 28 - page 17, line 25 * * page 18, line 6 - page 19, line 24 * * figures 2A,4,5B *	1,5-7, 12,16,17	

TECHNICAL FIELDS SEARCHED (Int.Cl.7)
H04Q
H04L

The present search report has been drawn up for all claims

Place of search	Date of completion of the search	Examiner
THE HAGUE	6 September 2000	Barel, C

CATEGORY OF CITED DOCUMENTS

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ANNEX TO THE EUROPEAN SEARCH REPORT ON EUROPEAN PATENT APPLICATION NO. EP 00 10 7329

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06-09-2000

Patent document cited in search report	Publication date	Patent family member(s)	Publication date
WO 9719525 A	29-05-1997	US 5729542 A EP 0804834 A	17-03-1998 05-11-1997
WO 9944379 A	02-09-1999	AU 2754899 A	15-09-1999

EPO FORM P0459

For more details about this annex : see Official Journal of the European Patent Office, No. 12/82

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(71) Applicant: NORTHERN TELECOM LIMITED
Montreal, Quebec H2Y 3Y4 (CA)

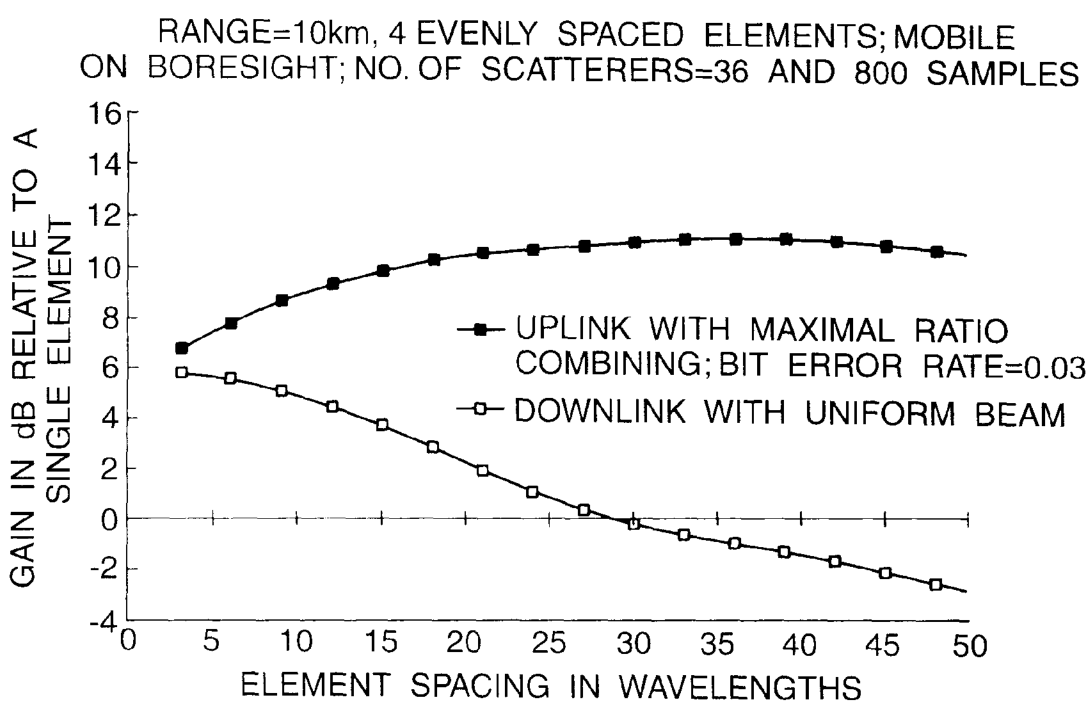
(72) Inventor: Smith, Martin Stevens
Chelmsford, Essex CM1 4XQ (GB)

(74) Representative: Dennis, Mark Charles et al
Nortel Patents,
London Road
Harlow, Essex CM17 9NA (GB)

(54) **An antenna downlink beamsteering arrangement**

(57) A base station arrangement including an antenna array is disclosed, wherein the uplink signals are weighted with complex array weights and the downlink signals are steered, wherein the downlink signals are steered using data from directional information derived from the uplink signals. A method of operation is also disclosed.

Fig.2.



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Description

This invention relates to cellular radio communication systems and in particular relates to an antenna downlink beamsteering arrangement.

Cellular radio systems are currently in widespread use throughout the world providing telecommunications to mobile users. In order to meet the capacity demand, within the available frequency band allocation, cellular radio systems divide a geographic area to be covered into cells. At the centre of each cell, there is a base station through which the mobile stations communicate, each base station typically being equipped with antenna arrays arranged sectors. Configurations of three or six sectors (sub-cells) are often employed, where the higher gain of correspondingly narrower beamwidth antennas improve the uplink from the lower power mobiles. The distance between the cells is determined such that co-channel interference is maintained at a tolerable level.

Obstacles in a signal path, such as buildings in built-up areas and hills in rural areas, act as signal scatterers and can cause signalling problems. These scattered signals interact and their resultant signal at a receiving antenna is subject to deep and rapid fading and the signal envelope often follows a Rayleigh distribution over short distances, especially in heavily cluttered regions. A receiver moving through this spatially varying field experiences a fading rate which is proportional to its speed and the frequency of the transmission. Since the various components arrive from different directions, there is also a Doppler spread in the received spectrum.

When a new cellular radio system is initially deployed, operators are often interested in maximising the uplink (mobile to base station) and downlink (base station to mobile station) range. The ranges in many systems are uplink limited due to the relatively low transmitted power levels of hand portable mobile stations. Any increase in range means that fewer cells are required to cover a given geographic area, hence reducing the number of base stations and associated infrastructure costs.

The range of the link, either the uplink or the downlink, can be controlled principally in two different ways: by adjusting either the power of the transmitter or the gain at the receiver. On the downlink the most obvious way of increasing the range is to increase the power of the base station transmitter. To balance the link the range of the uplink must also be increased by an equivalent amount. The output power of a transmitter on a mobile, however, is constrained to quite a low level to meet national regulations, which vary on a country to country basis. Accordingly the receive gain at the base station must be increased.

The principal method of improving the receive system gain and to reduce the effect of fading is to include some form of diversity gain in addition to the receive antenna gain. The object of a diverse system is to provide the receiver with more than one path, with the paths being differentiated from each other by some means, e.g. space, angle, frequency or polarisation. The use of these additional paths by the receiver provides the diversity gain. The amount of gain achieved depends upon the type of diversity, number of paths, and method of combination.

This invention is concerned with spatially diverse systems and in particular seeks to provide an arrangement wherein downlink performance is improved.

Cellular radio base stations frequently use two antennas for diversity reception on the uplink, spaced by many (e.g. 20) wavelengths. This large spacing is required because the angular spread of the incoming signals is narrow. This can be represented as a ring of scatterers around a mobile user who is transmitting to a base station otherwise known as the uplink path and such an arrangement is shown in Figure 1. For example the radius of scatterers may be 50 to 100 metres, and the range to the base station may be up to 10 km, resulting in a narrow angular spread. A large antenna spacing is required at the basestation to provide decorrelated fading , which can be calculated from the Fourier transform relationship between antenna array aperture and angular width (a large aperture in wavelengths provides a narrow beam).

In order to improve wanted signals and discriminate against interfering signals, antennas are being developed which utilise an array of antenna elements at the base station, allied with an "intelligent" beamformer. One such technique is to use a multichannel maximal ratio combiner on reception at the base station array. This operates by weighting the array signals s_i (i=1 to N, where N = the number of elements in the array) with their complex conjugates s_i^* (assuming equal noise powers on each channel) and summing to give:

$$S = \sum_{i=1}^{N} s_i^* s_i = \sum_{i=1}^{N} |s_i|^2.$$

For a N element array, this provides both array gain (approximately a factor N in power) and diversity gain, the latter only if at least some of the array elements are widely spaced. Thus a factor N improvement in mean signal level can be achieved, allowing extended range or lower mobile transmit power. The array provides narrower beams than a single antenna element, and hence also provides better protection against interference, improving carrier to interference ratios and hence allowing higher capacity systems by reducing re-use factors.

The limitation of the above is that the improvements are only for the uplink, and not for the downlink (base station transmit to the mobile). The present invention seeks to provide an improved downlink signal.

A standard feature of a number of cellular radio systems is that the sets of uplink and downlink frequencies

are separated into two distinct bands spaced by a guard band, for example 1800 - 1850 MHz (uplink) and 1900 - 1950 MHz (downlink). Up- and down- link frequencies are then paired off, e.g. 1800 with 1900, 1850 with 1950. There is therefore a significant change of frequency (e. g. 5%) between up and down links. There is consequently no correlation for the fast fading (as the mobile moves) between up and down links.

In accordance with the present invention, there is provided a base station arrangement including an antenna array, wherein the uplink signals are weighted with complex array weights and wherein the downlink signals are steered using directional information derived from the uplink signals.

In accordance with another aspect of the present invention, common array elements are used for the uplink and downlink signals. Alternatively, only some of the antenna elements are employed for both the uplink and downlink signals. Separate arrays can be used for the up and down links, and in particular it may be preferable to have a closely spaced array for the downlink, with a less closely spaced array for the uplink.

In accordance with a still further aspect of the invention, there is provided a base station arrangement, wherein the antennas are arranged in two groups per facet, wherein a first group comprises a plurality of antenna arrays and a second group comprises a single antenna array. Alternatively, both group could comprise a plurality of antenna arrays.

In accordance with a still further aspect of the invention, there is provided a method of operating a base station arrangement, wherein incoming signals from a mobile radio are weighted with complex array weights, deriving directional information from these signals and applying the directional information to the downlink signals whereby a downlink beam is steered towards the mobile.

The method of combining the uplink signal can be performed by the use of maximal ratio combining, with the method of combining the downlink signal employing standard beam weights. Non-uniform array spacings can be used.

The present invention thus resides in the use of complex array weights for the uplink signals, deriving directional information from the uplink signals and using this data to steer the downlink beam.

In order that the invention may be more fully understood, reference will now be made to the figure as shown in the accompanying drawing sheets, wherein:

Figure 1 shows a downlink signal scattering model ;
Figure 2 is a graph detailing uplink and downlink gain versus antenna element spacing for a 4-element antenna array, with a mobile at broadside; and
Figure 3 is a graph detailing uplink and downlink gain versus antenna element spacing for a 4-element antenna array, with a mobile at 30° from broadside.

Figure 2 shows the array gain for a four element array, where maximal ratio combining weights are used for the uplink and a standard beam (e.g. uniform amplitude array weights) are used for the downlink. The gain is shown as a function of array inter-element spacing. This figure shows gain averaged through the fast fading, and are for the case of a mobile positioned "broadside" to the array. The uplink gain rises above 6 dB (N=4) due to diversity gain (this part is dependent on the error rate). No diversity gain occurs on the downlink, as standard beam weights are used. Significant array gain is available on the downlink, provided the array spacing is not too large. It is then possible to select an array spacing such that array gain and significant diversity gain are available on the uplink, and there is still significant array gain for the downlink, for example with an array spacing of about 10 wavelengths for this scenario.

Figure 3 shows the corresponding results for the case where the mobile position is moved to 30 degrees from broadside, and direction finding (d.f.) using the uplink signals has been employed to steer the downlink beam towards the mobile and its ring of scatterers. The resulting curve is similar to the broadside case, apart from a factor to allow for the projected aperture of the array.

Two possible uplink/downlink scenarios arise from these results: Common array elements can be used with complex weights (e.g. maximal ratio combining weights) for the uplink and standard beam weights (uniform or tapered amplitude, phase slope to steer the beam) for the downlink. Alternatively, separate arrays can be used for up and down links, for example a closely spaced array can be employed for the downlink, to provide the maximum downlink gain (the left portion of the graphs in Figures 2 and 3), with a less closely spaced array being employed for the uplink, to provide maximum spatial diversity (the centre-right portions of the graphs in Figures 2 and 3). A combination of these two concepts is also possible, for example, if some elements are shared and non-uniform array spacings are used. Thus, complex array weights are employed for the uplink, the downlink beam is steered, with directional information being derived from the uplink signals.

There are various possible methods for deriving directional information from the uplink signals. One example is to use an array with a first group of closely spaced elements (< 1λ), plus one or more antenna elements which are spaced from the first group of elements and can be considered as "out-lier" elements with a wide spacing to the close spaced group, to achieve good spatial diversity gain for the uplink. The out-lier elements may comprise a single linear array or comprise a second group of elements, conveniently the same type of array as the first group whereby uniformity of componentry may be maintained and reduce costs of manufacture and ease installation.

The first group of elements (and second if of a smilar configuration) can be connected to a multiple beam

former, such as a Butler matrix, which forms simultaneous multiple beams spanning the sector of interest. By detecting the relative amplitudes in the multiple beams, the angle of arrival of the uplink signal can be deduced, and this information used to derive the necessary phase slope to be applied to the close spaced array elements for the downlink signal. Uplink maximal ratio combining can be performed on the complex beam outputs plus the outlier element(s) output(s).

Since direction finding is facilitated with an array containing both small and large spacings, this array configuration is also usefully incorporated for the uplink.

There are four antenna columns on a typical cellular base station facet: on the uplink all four antenna columns are used and maximal ratio combining is carried out; on the downlink, rather than combining the outputs through four transmitters, the signals are fed through one antenna. The combining advantages are lost on the downlink since the antennas of a whole array are employed for each frequency . The present invention allows the burden of combining to be shared, where there is an out-lier, whereby spatial diversity is obtained by spacing the antenna groups spaced apart. Signals do not have to be put through the transceiver transmitters of only one group of anennas of one facet: instead the signals can be split between the grpoups of antennas of the facet. This eases the combining load imposed on the antennas and beamformers. A further advcantage lies in the reduced visual impact of a base station. Whilst there are two antenna groups per sector, which increases the number of elements liable to sreate a visual impact, the size of the antenna groups can be reduced whereby a smaller visual impact is created, provided that the antenna groups are sufficiently widely spaced apart.

Claims

1. A base station arrangement comprising an antenna array, wherein the uplink signals are weighted with complex array weights and wherein the downlink signals are steered using directional information derived from the uplink signals.

2. An arrangement according to claim 1 wherein common array elements are used for the uplink and downlink signals.

3. An arrangement according to claim 1 wherein some antenna elements are employed for both the uplink and downlink signals.

4. An arrangement according to claim 1 wherein separate arrays are used for up and down links,

5. An arrangement according to claim 4 wherein a closely spaced array is employed for the downlink, with a less closely spaced array being employed for the uplink.

6. An arrangement according to claim 1 or 2, wherein the method of combining the uplink signal is maximal ratio combining.

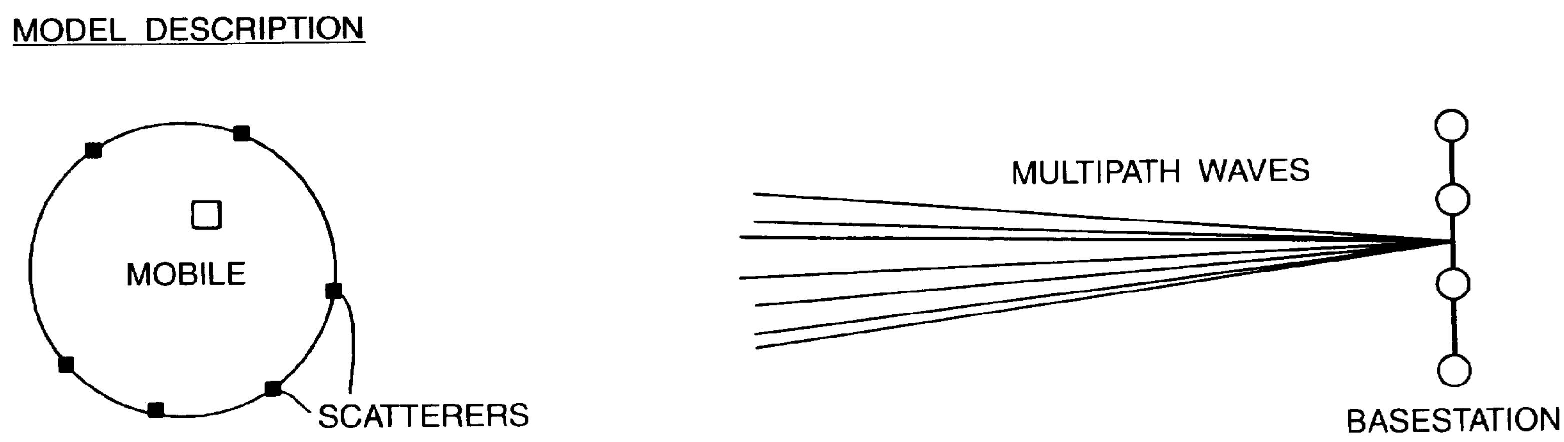
7. An arrangement according to claim 1 or 2, wherein the method of combining the downlink signal employs standard beam weights.

8. An arrangement according to any one of claims 1 to 7, wherein the antennas are arranged in two groups per facet, wherein a first group comprises a plurality of antenna arrays and a second group comprising a single antenna array or a plurality of antennas.

9. An arrangement according to any one of claims 1 to 7, wherein non-uniform array spacings are used.

10. A method of operating a base station arrangement, wherein incoming signals from a mobile radio are weighted with complex array weights, deriving directional information from these signals and applying the directional information to the downlink signals whereby a downlink beam is steered towards the mobile radio.

Fig.1.



A SIGNAL TRANSMITTED FROM THE MOBILE REACHES THE BASESTATION HAVING TRAVELLED VIA A NUMBER OF PATHS WHICH EXIST AS A RESULT OF SCATTERING FROM OBSTACLES RANDOMLY DISTRIBUTED ON A CIRCLE SURROUNDING THE MOBILE

Fig.2.

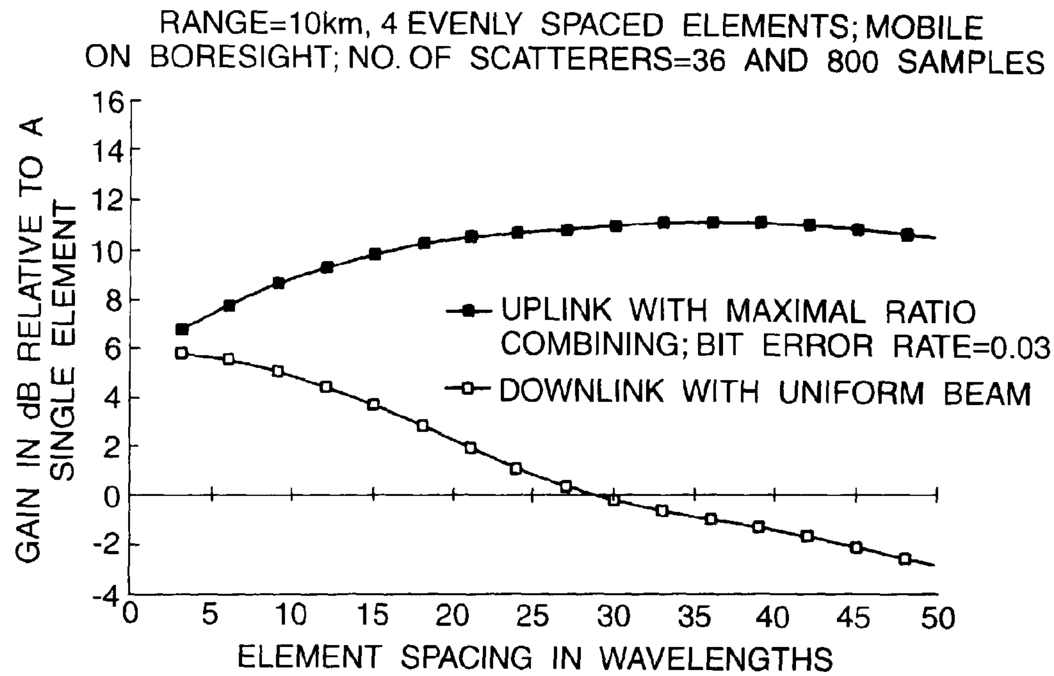
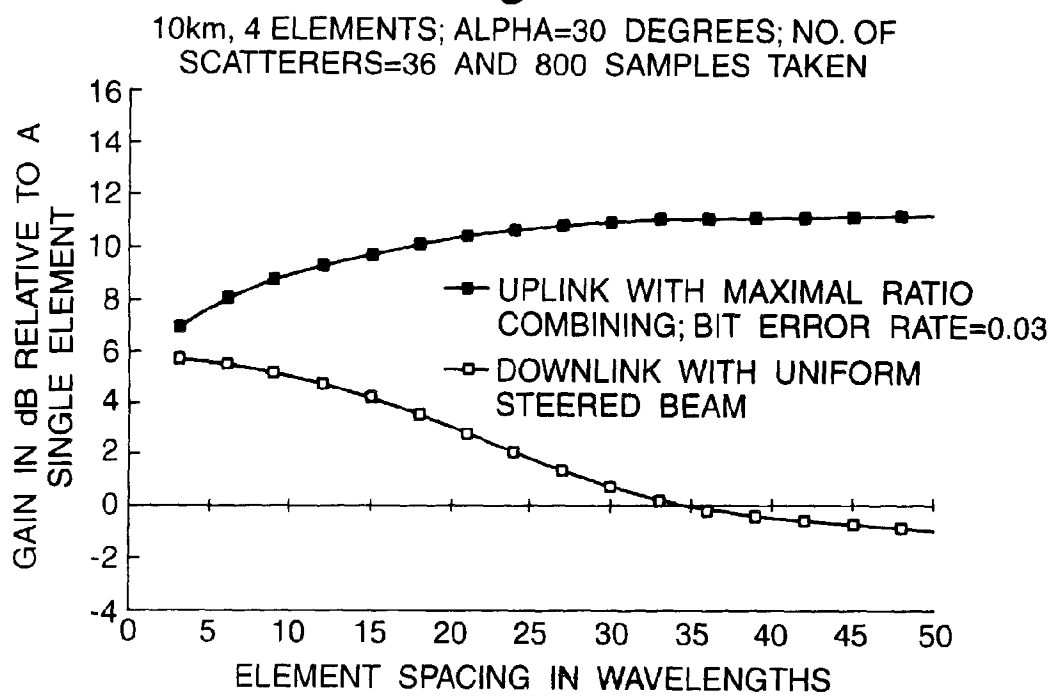

RANGE=10km, 4 EVENLY SPACED ELEMENTS; MOBILE ON BORESIGHT; NO. OF SCATTERERS=36 AND 800 SAMPLES
GAIN IN dB RELATIVE TO A SINGLE ELEMENT
ELEMENT SPACING IN WAVELENGTHS
UPLINK WITH MAXIMAL RATIO COMBINING; BIT ERROR RATE=0.03
DOWNLINK WITH UNIFORM BEAM
16
14
12
10
8
6
4
2
0
-2
-4
0
5
10
15
20
25
30
35
40
45
50

Fig.3.


10km, 4 ELEMENTS; ALPHA=30 DEGREES; NO. OF SCATTERERS=36 AND 800 SAMPLES TAKEN
GAIN IN dB RELATIVE TO A SINGLE ELEMENT
ELEMENT SPACING IN WAVELENGTHS
UPLINK WITH MAXIMAL RATIO COMBINING; BIT ERROR RATE=0.03
DOWNLINK WITH UNIFORM STEERED BEAM
16
14
12
10
8
6
4
2
0
-2
-4
0
5
10
15
20
25
30
35
40
45
50

European Patent Office

EUROPEAN SEARCH REPORT

Application Number
EP 96 30 4416

DOCUMENTS CONSIDERED TO BE RELEVANT

Category	Citation of document with indication, where appropriate, of relevant passages	Relevant to claim	CLASSIFICATION OF THE APPLICATION (Int.Cl.6)
X Y A	WO-A-94 09568 (E-SYSTEMS) 28 April 1994 * abstract * * claim 17; figures 1,2,5,6 *	1,2,7,10 9 3-6	H01Q3/26 H01Q25/00
X	US-A-5 260 968 (GARDNER ET AL.) 9 November 1993 * abstract; claims 1,9,13,16,21; figures 6,7 *	1,4,7,10	
X	EP-A-0 595 247 (ATR OPTICAL AND RADIO COMMUNICATIONS RESEARCH) 4 May 1994 * abstract; claims 1,5; figures 1,2 *	1,10	
X	GB-A-2 266 998 (MOTOROLA) 17 November 1993 * abstract; figures 1,2 *	1,10	
Y	US-A-4 104 641 (UNZ) 1 August 1978 * abstract; figure *	9	
Y	EP-A-0 374 008 (THOMSON-CSF) 20 June 1990 * abstract; figures 1,2 *	9	TECHNICAL FIELDS SEARCHED (Int.Cl.6) H01Q

The present search report has been drawn up for all claims

Place of search	Date of completion of the search	Examiner
THE HAGUE	30 August 1996	Angrabeit, F

CATEGORY OF CITED DOCUMENTS

X : particularly relevant if taken alone
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L : document cited for other reasons

& : member of the same patent family, corresponding document

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(71) Applicant: DIGI-MEDIA VISION LIMITED
West Drayton, Middlesex UB7 0DQ (GB)

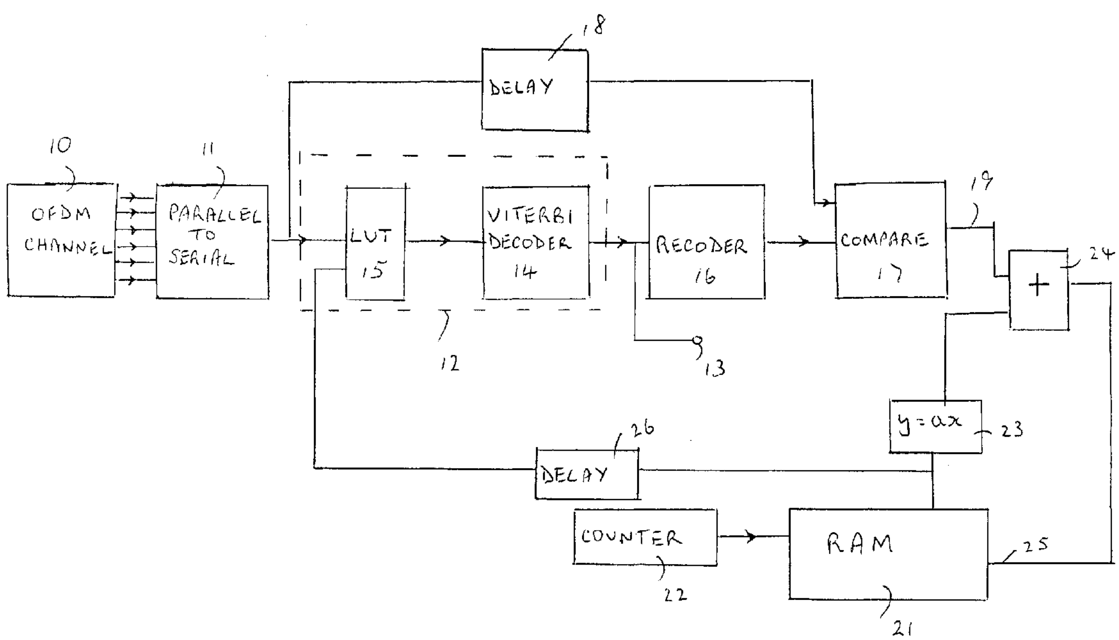
(72) Inventors:
- Gledhill, Jeff
Eastleigh, Hampshire (GB)
- Anikhindhi, Santosh
Chandlers Ford, Hampshire SO53 3PQ (GB)

(74) Representative: Brown, Angela Mary
Gamma House,
Enterprise Road
Chilworth, Hampshire SO16 7NS (GB)

(54) Method and apparatus for decoding orthogonal frequency division multiplexed (OFDM) carriers

(57) The invention relates to decoding symbols in the form of modulated carriers which are encoded using orthogonal frequency division multiplexing.

The modulated carriers are supplied to a decoder which is preferably a Viterbi decoder. The decoder symbols are then recoded and passed to a comparator where they are compared with the incoming modulated carriers to locate errors between each recoded symbol and the corresponding modulated carrier symbol. The error rate for each carrier is derived by a counter and the indications of error rate and the modulated carriers are used to address entries in a look up table from which a decoder decodes the symbols represented by the modulated carriers.



EP 0 772 329 A1

Description

The present invention relates to the decoding of symbols transmitted to a receiver which receives the symbols in the form of modulated carriers encoded using orthogonal frequency division multiplexing. At the receiver, the encoded symbols are decoded by a suitable decoder such as a Viterbi decoder.

A problem in decoding the modulated carriers occurs if the carriers are subject to fading or interference. The problem is particularly acute if fading or interference affects the carriers over a narrow band of frequency since, although there may a negligible effect on the majority of carriers, it causes a large effect on a few carriers. The problem may be the result of frequency selective fading caused by multipath interference or may be the result of narrow band interference caused by analog television signals.

The use of patterns of erasures to increase the rate of error correction of Viterbi decoded convolutional codes is already known. This technique, known as puncturing, uses short repetitive sequences of erasures and achieves very good performance. Simulations using random patterns of erasures for puncturing give results that are only slightly inferior to the best known repetitive sequences.

Rather than simply marking a symbol as erased or not erased, an error probability can be assigned to the symbol to represent information about the reliability of the received symbol. This technique is widely used in the so-called soft decision Viterbi decoder and in its standard from the information on the reliability of the symbol is derived solely from the quality of the received symbol, usually as a function of its Euclidean distance from a decision threshold.

In an extended form of the Viterbi decoder, soft decision information is derived from the state of the channel. One technique is to determine the signal to noise ratio for each carrier and to use this information to derive a measure of the error probability for the symbols assigned to that carrier. One method for determining this error probability is to use a null symbol, during which there is an interruption in the transmission, and to allow the receiver to measure the spectral response of the noise and interference and thereby infer the interference power and hence error probability for each carrier.

A problem with the use of null symbols is that the data rate of the transmission is degraded by the need to accommodate the null symbols.

According to the present invention there is provided a method of decoding incoming modulated carriers encoded using orthogonal frequency division multiplexing and representing a stream of encoded symbols, the method comprising the steps of; supplying the modulated carriers to a decoding means effective to decode the symbols represented thereby, recoding the decoded symbols, comparing the incoming modulated carriers with the recoded symbols to locate errors between each recoded symbol and the corresponding modulated carrier symbol, deriving an indication of the error rate for each carrier and, applying the indication of error rate to the decoding means to effect decoding of each carrier symbol by reference to the modulation of the carrier combined with the indication of error rate for the carrier.

According to the present invention there is now provided apparatus for decoding incoming modulated carriers encoded using orthogonal frequency division multiplexing and representing a stream of symbols, the apparatus comprising; decoding means having an input to receive the modulated carriers and being effective to decode the symbols represented thereby, a recoder to recode the decoded symbols from the decoding means, a comparator to compare the incoming modulated carriers with the recoded symbols and to signal the location of errors between each recoded symbol and the corresponding modulated carrier symbol, and means to derive from the comparator an indication of the error rate for each carrier, the decoding means having an input to receive the indication of error rate for each carrier and being effective to decode each carrier symbol by reference to the modulation of the carrier and by reference to the indicated error rate for the carrier.

Preferably, the decoding means comprise a look up table, entries to which are addressed by the carriers and the indications of error rate, and a decoder effective to decode the symbols represented by the modulated carriers from the entries addressed in the look up table.

The invention will now be described, by way of example, with reference to the accompanying drawing which shows a system embodying the present invention for decoding modulated carriers encoded using orthogonal frequency division encoding.

A channel 10 supplies in parallel a plurality of modulated carriers encoded using orthogonal frequency division multiplexing. The carriers are received by a de-interleaver 11 which converts the carriers from parallel to serial format to represent a stream of encoded symbols. A decoding means 12 is connected to receive the stream of symbols and is effective to decode the symbols for supply to an output terminal 13. The decoding means consist of a Viterbi decoder 14 and means to store a look up table 15. The purpose of the look up table will be explained.

The output from the Viterbi decoder 14 is supplied to a recoder 16 which recodes the symbols decoded by the Viterbi decoder and supplies the recoded symbols to one input to a comparator 17. Another input to the comparator 17 is supplied with the incoming carriers representing the input stream of encoded symbols by way of a delay means 18. The delay means 18 delay the stream of encoded symbols passed therethrough by a delay period sufficient to match the delay through the look up table means 15, the Viterbi decoder 14 and the recoder 16. As a result, each symbol in the incoming stream from the de-interleaver 11 is passed through the delay means 18 to be compared with the recoded ver-

sion of the same symbol supplied by the recoder 16. The comparator 17 is effective to signal to an output terminal 19 each error arising from a difference in the coding of the symbol supplied by way of the delay means and the coding of the corresponding symbol supplied by the recoder 16.

The error signals at the terminal 19 are passed to a bit error counter. The bit error counter includes a RAM 21 which stores a series of n bit numbers associated respectively with the bit positions in a received symbol. The numbers in the RAM 21 are addressed sequentially by a counter 22 which steps through the RAM addresses so that all the n bit numbers are addressed over a time interval of a received symbol. The RAM has an output connected to a multiplier 23 which supplies one input of an adder 24. The other input to the adder 24 is the output from the comparator 17. The adder supplies an input 25 of the RAM 21.

Each number addressed in the RAM is multiplied by a factor a in the multiplier 23, the value of a being less than 1. The output from the multiplier 23 is returned to the same RAM address by way of the adder 24. In the event of no error, the value returned to the RAM by way of the adder 24 is simply the number addressed in the RAM multiplied by the value a. The result is stored back in the RAM 21.

In the event that the comparator 17 detects any error, the number addressed in the RAM is multiplied by the value a and the result is added, in the adder 24, to a number b before being stored back in the RAM. It can be shown that for a long term error probability p, the output from the RAM 21 is given by the expression;

$$\frac{b}{(1 - ap)} -b$$

The output from the RAM 21 is supplied through a delay means 26 to the look up table 15. The delay means 26 impose a delay equal to the length of one received symbol so that the numbers from the RAM 21 correspond in time to the bit positions of the next received symbol in the stream of encoded symbols from the de-interleaver 11.

The look up table 15 combines the encoded symbol with the bit error rate information form the RAM 21 to generate the input to the Viterbi decoder 14. The combined input to the Viterbi decoder 17 is exactly the information required by the decoder 17 to optimally decode the received stream of encoded signals. The estimation of the performance of the channel 10 is thus recovered directly from the received symbols rather than by inference from other channel measurements and without the use of special signals such as a null symbol.

Claims

1. A method of decoding modulated carriers encoded using orthogonal frequency division multiplexing and representing the stream of encoded symbols, the method comprising the steps of: supplying modulated carriers to a decoding means effective to decode the symbols represented thereby, recoding the decoded symbols, comparing the modulated carriers with the recoded symbols to locate errors between each recoded symbol and the corresponding modulated carrier symbol, deriving an indication of the error rate for each carrier and, applying the indication of error rate to the decoding means to effect decoding of each carrier symbol by reference to the modulation of the carrier combined with the indication of error rate for the carrier.

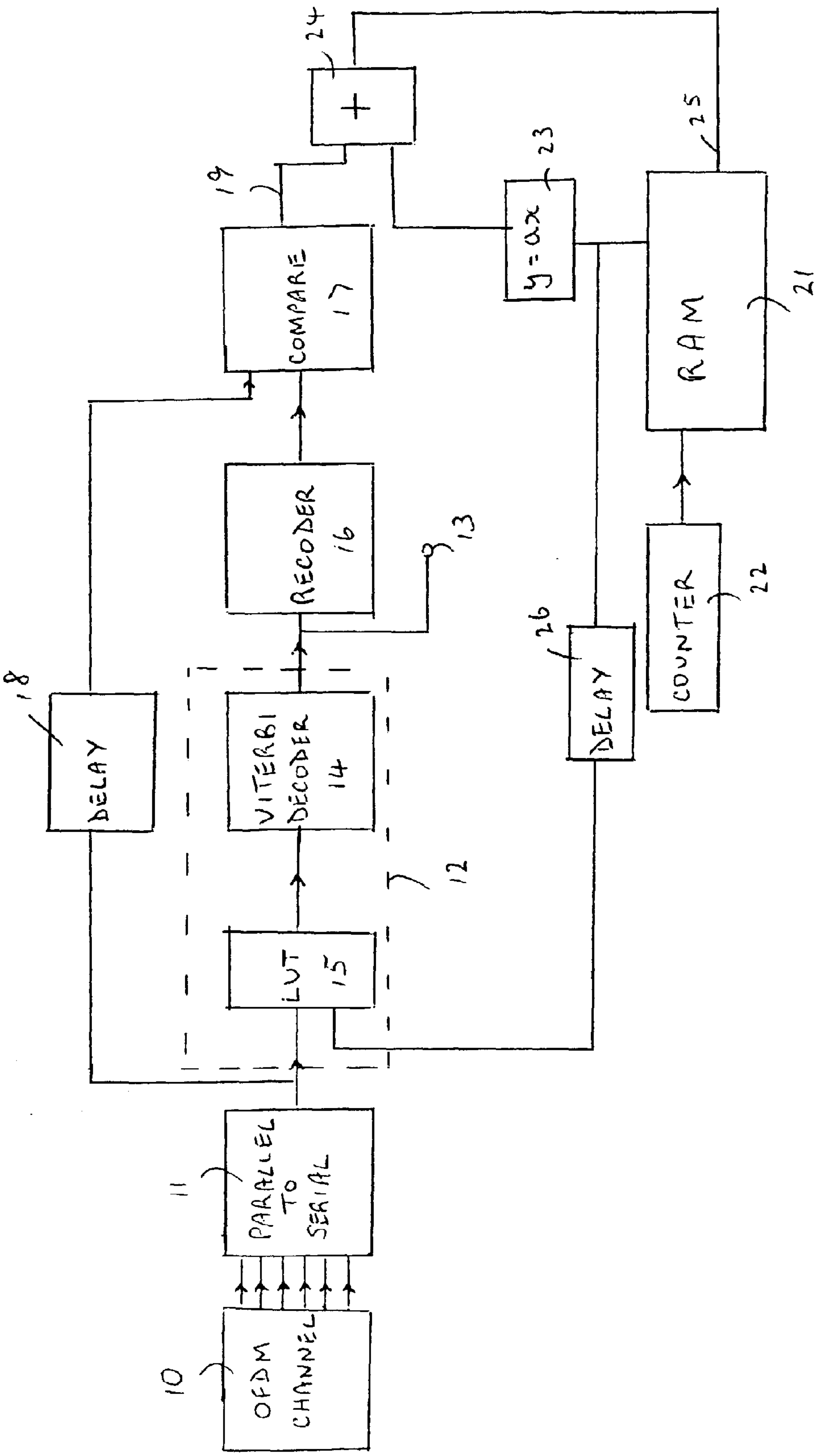
2. A method as claimed in claim 1, wherein the indications of error rate and the modulated carriers are used to address entries in a look up table from which a decoder decodes the symbols represented by the modulated carriers.

3. A method as claimed in claim 1 or 2, wherein the decoding means comprise a Viterbi decoder.

4. Apparatus for decoding modulated carriers encoded using orthogonal frequency division multiplexing, the apparatus comprising, decoding means having an input to receive the modulated carriers and being effective to decode the symbols represented thereby, a recoder to recode the symbols from the decoding means, a comparator to compare the modulated carriers with the recoded symbols and to signal the location of errors between each recoded symbol and the corresponding modulated carrier symbol and, means to derive from the comparator an indication of the error rate for each carrier, the decoding means having an input to receive the indication of error rate for each carrier and being effective to decode each carrier symbol by reference to the modulation of the carrier and by reference to the indicated error rate for the carrier.

5. Apparatus as claimed in claim 4, wherein the decoding means comprise a look up table, entries to which are addressed by the carriers and the indications of error rate, the decoding means comprising a decoder effective to decode the symbols represented by the modulated carriers from the entries addressed in the look up table.

6. Apparatus as claimed in claim 4 or 5, wherein the decoding means comprise a Viterbi decoder.


OFDM CHANNEL
10
PARALLEL TO SERIAL
11
LUT
15
VITERBI DECODER
14
12
13
RECODER
16
COMPARE
17
DELAY
18
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+
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y = ax
23
RAM
21
25
COUNTER
22
DELAY
26

European Patent Office

EUROPEAN SEARCH REPORT

Application Number
EP 96 30 7658

DOCUMENTS CONSIDERED TO BE RELEVANT

Category	Citation of document with indication, where appropriate, of relevant passages	Relevant to claim	CLASSIFICATION OF THE APPLICATION (Int.Cl.6)
X	IBC 95. INTERNATIONAL BROADCASTING CONVENTION (CONF. PUBL. NO. 413), 14 - 18 September 1995, ISBN 0-85296-644-X, 1995, LONDON, UK, IEE, UK, pages 122-128, XP000617513 MIGNONE V. ET AL.: "CD3-OFDM: a new channel estimation method to improve the spectrum efficiency in digital terrestrial television systems"	1,3,4,6	H04L27/26 H04L5/06
A	* section 3 * * figure 3 *	2,5	
A	US 5 278 871 A (RASKY ET AL.) 11 January 1994 * abstract; figure 1 *	1,4	
A	EBU REVIEW- TECHNICAL, no. 224, 1 August 1987, pages 168-190, XP000560523 LASSALLE R. ET AL.: "PRINCIPLES OF MODULATION AND CHANNEL CODING FOR DIGITAL BROADCASTING FOR MOBILE RECEIVERS" * abstract * * section 4 *	1-6	TECHNICAL FIELDS SEARCHED (Int.Cl.6) H04L
A	PATENT ABSTRACTS OF JAPAN vol. 012, no. 420 (E-679), 8 November 1988 & JP 63 156444 A (MITSUBISHI), 29 June 1988, * abstract *	1,4	

The present search report has been drawn up for all claims

Place of search	Date of completion of the search	Examiner
THE HAGUE	14 February 1997	Ghigliotti, L

CATEGORY OF CITED DOCUMENTS

X : particularly relevant if taken alone
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(71) Applicant: **LUCENT TECHNOLOGIES INC.**
Murray Hill, New Jersey 07974-0636 (US)

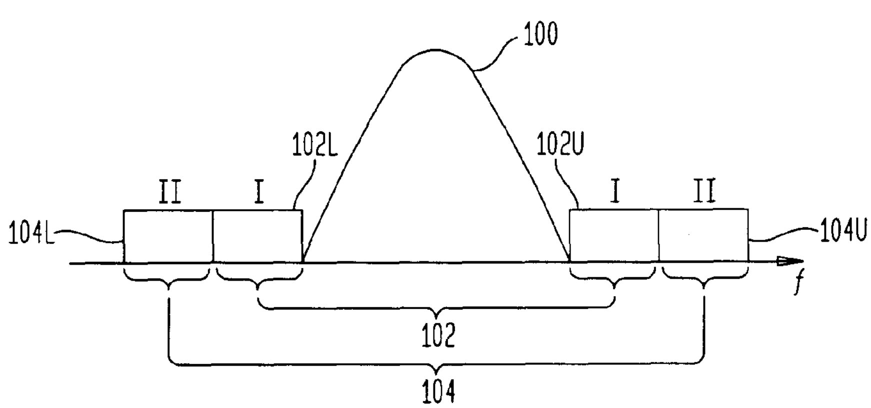
(72) Inventors:
- **Sinha, Deepen**
Chatham, New Jersey 07928 (US)
- **Sundberg, Carl-Erik Wilhelm**
Chatham, New Jersey 07928 (US)

(74) Representative:
Watts, Christopher Malcolm Kelway, Dr. et al
Lucent Technologies (UK) Ltd,
5 Mornington Road
Woodford Green Essex, IG8 0TU (GB)

(54) Unequal error protection for digital broadcasting using channel classification

(57) The invention provides methods and apparatus for processing information, e.g., audio, video or image information, for transmission in a communication system. In an illustrative embodiment, interference characteristics are determined for a set of *n* channels to be used to transmit audio information bits, where *n* is greater than or equal to two. The audio information bits are separated into n classes based on error sensitivity, for example, the impact of errors in particular audio data bits on perceived quality of an audio signal reconstructed from the transmission. The classes of bits are then assigned to the *n* channels such that the classes of bits having the greatest error sensitivity are transmitted over the channels which are the least susceptible to interference. The interference characteristics associated with the *n* channels can be determined by, for example, measuring interference levels for one or more of the channels, or obtaining information regarding known interference levels for one or more of the channels. The channels may correspond to different frequency bands, time slots, code division slots or any other type of channels. The invention can provide UEP for different classes of audio information bits even in cases in which the same convolutional code, or the same complementary punctured pair convolutional (CPPC) code pair, is used to encode the classes. The assignment of the classes of bits to the channels, as well as the characteristics of the classes and the channels, may be fixed or dynamic.

FIG. 1



EP 0 991 221 A2

Description

Field of the Invention

[0001] The present invention relates generally to digital audio broadcasting (DAB) and other techniques for transmitting information, and more particularly to techniques for providing unequal error protection (UEP) for different classes of audio, video, image or other information bits encoded in a source coding device.

Background of the Invention

[0002] Most source coded bit streams exhibit unequal sensitivity to bit errors. For example, certain source bits can be much more sensitive to transmission errors than others. Moreover, errors in certain bits, such as control bits, may lead to severe error propagation and a corresponding degradation in reconstructed signal quality. Such error propagation can occur, for example, in the output audio bits of an audio coder due to the use of control bits for codebook information, frame size information, synchronization information and so on. The perceptual audio coder (PAC) described in D. Sinha, J.D. Johnston, S. Dorward and S.R. Quackenbush, "The Perceptual Audio Coder," in Digital Audio, Section 42, pp. 42-1 to 42-18, CRC Press, 1998, which is incorporated by reference herein, attempts to minimize the bit rate requirements for the storage and/or transmission of digital audio data by the application of sophisticated hearing models and signal processing techniques. In the absence of channel errors, a PAC is able to achieve near stereo compact disk (CD) audio quality at a rate of approximately 128 kbps. At a lower bit rate of 96 kbps, the resulting quality is still fairly close to that of CD audio for many important types of audio material.

[0003] The rate of 96 kbps is particularly attractive for FM band transmission applications such as in-band digital audio broadcasting (DAB) systems, which are also known as hybrid in-band on-channel (HIBOC), all-digital IBOC and in-band adjacent channel (IBAC)/in-band reserve channel (IBRC) DAB systems. There is also a similar effort underway to provide digital audio broadcasting at lower audio bit rates in the AM band. For these AM systems, audio bit rates of about 32 to 48 kbps are being considered for daytime transmission and about 16 kbps for nighttime transmission. Higher audio bit rates, greater than about 128 kbps, are being used in multiple channel DAB systems. The transmission channels in the above-noted DAB systems tend to be severely bandlimited and noise limited at the edge of a coverage area. For mobile receivers, fading is also a severe problem. It is therefore particularly important in these and other applications to design an error protection technique that is closely matched to the error sensitivity of the various bits in the compressed audio bit stream.

[0004] PACs and other audio coding devices incorporating similar compression techniques are inherently packet-oriented, i.e., audio information for a fixed interval (frame) of time is represented by a variable bit length packet. Each packet includes certain control information followed by a quantized spectral/subband description of the audio frame. For stereo signals, the packet may contain the spectral description of two or more audio channels separately or differentially, as a center channel and side channels (e.g., a left channel and a right channel). Different portions of a given packet can therefore exhibit varying sensitivity to transmission errors. For example, corrupted control information leads to loss of synchronization and possible propagation of errors. On the other hand, the spectral components contain certain interframe and/or interchannel redundancy which can be exploited in an error mitigation algorithm incorporated in a PAC codec. Even in the absence of such redundancy, the transmission errors in different audio components have varying perceptual implications. For example, loss of stereo separation is far less annoying to a listener than spectral distortion in the mid-frequency range in the center channel.

[0005] Unequal error protection (UEP) techniques are designed to match error protection capability with sensitivity to transmission errors, such that the most important bits are provided with the highest level of error protection, while less important bits are provided with a lesser level or levels of error protection. A conventional two-level UEP technique for use in DAB applications is described in N.S. Jayant and E.Y. Chen, "Audio Compression: Technology and Applications," AT&T Technical Journal, pp. 23-34, Vol. 74, No. 2, March-April 1995. In this technique, which is based on a Reed-Solomon (RS) code, the control information is protected more robustly since it is not possible to use error mitigation on the non-redundant control information. In fact, the proper operation of the error mitigation algorithm used in a PAC codec is itself dependent upon reliable control information. All of the non-control spectral information in this technique is protected using a uniform level of error protection.

[0006] EP-A-0936772 discloses techniques for providing UEP of a PAC bitstream by classifying the bits in different categories of error sensitivity. These classes were then matched to a suitable level of error protection to minimize the overall impact of errors, i.e., the most sensitive bits are more protected than the others. Certain of the UEP techniques described in the above-cited application generally provide improvements without regard to the type of channel, and the channel noise is typically assumed to be averaged over time and frequency by interleaving in both time and frequency for each channel code class. Thus, a UEP technique with a more powerful channel code properly matched to the most sensitive source bits always outperforms the corresponding equal error protection (EEP) technique. However,

determining the channel codes for such UEP scenarios is often a nontrivial problem, particularly in the case of determining single sideband complementary punctured-pair convolutional codes (CPPC) codes for HIBOC applications. Therefore, although the techniques in the above-cited application provide considerable improvement over prior approaches to UEP for digital audio, further improvements are needed for certain implementations, such as the above-noted HIBOC systems and other similar systems.

Summary of the Invention

[0007] The present invention provides methods and apparatus for implementing UEP for a source coded bit stream such as that generated by a perceptual audio coder (PAC). In an illustrative embodiment, interference characteristics are determined for a set of *n* channels to be used to transmit audio information bits, where *n* is greater than or equal to two. The audio information bits are separated into *n* classes based on error sensitivity, for example, the impact of errors in particular audio data bits on perceived quality of an audio signal reconstructed from the transmission. The classes of bits are then assigned to the *n* channels such that the classes of bits having the greatest error sensitivity are transmitted over the channels which are the least susceptible to interference. The interference characteristics associated with the *n* channels can be determined by, for example, measuring interference levels at different times and locations for one or more of the channels, or obtaining information regarding known interference levels for one or more of the channels. The channels may correspond to different frequency bands, time slots, code division slots or any other type of channels. The channel properties may also change with factors such as time and location within a coverage area.

[0008] In accordance with another aspect of the invention, the assignment of the classes of bits to the channels, as well as the characteristics of the classes and the channels, may be fixed or dynamic. For example, in applications in which the interference characteristics associated with one or more of the channels vary as a function of time, position within a coverage area, or other factors, the assignment of the classes of bits to the channels can be varied so as to ensure that the classes of bits having the greatest error sensitivity continue to be transmitted over the channels which are least susceptible to interference. As another example, amounts of channel resources used for particular classes of audio information bits can vary as a function of time.

[0009] The invention can provide UEP for different classes of information bits even in cases in which the same convolutional code, or the same CPPC code pair, is used to encode the classes, although different channel codes could also be used to encode the classes. The invention can be applied to other types of digital information, including, for example, video and image information. Moreover, the invention is applicable not only to perceptual coders but also to other types of source encoders using other compression techniques operating over a wide range of bit rates, and can be used with transmission channels other than radio broadcasting channels.

Brief Description of the Drawings

[0010] FIG. 1 illustrates a two-class frequency division unequal error protection (UEP) technique is accordance with the invention as applied to an exemplary hybrid in-band on-channel (HIBOC) digital audio broadcasting (DAB) system.

[0011] FIGS. 2 through 4 illustrate a number of possible alternative implementations of the two-class UEP technique of FIG. 1.

[0012] FIG. 5 is a block diagram of a communication system in which an n-class frequency division UEP technique is implemented in accordance with an illustrative embodiment of the invention.

Detailed Description of the Invention

[0013] The invention will be described below in conjunction with exemplary unequal error protection (UEP) techniques for use in the transmission of audio information bits, e.g., audio bits generated by an audio coder such as the perceptual audio coder (PAC) described in D. Sinha, J.D. Johnston, S. Dorward and S.R. Quackenbush, "The Perceptual Audio Coder," in Digital Audio, Section 42, pp. 42-1 to 42-18, CRC Press, 1998. It should be understood, however, that the UEP techniques of the invention may be applied to many other types of information, e.g., video or image information, and other types of coding devices. In addition, the invention may be utilized with a wide variety of different types of communication applications, including communications over the Internet and other computer networks, and over cellular multimedia, satellite, wireless cable, wireless local loop, high-speed wireless access and other types of communication systems. Although illustrated at least in part using frequency bands as channels, the invention may also be applied to many other types of channels, such as, for example, time slots, code division multiple access (CDMA) slots, and virtual connections in asynchronous transfer mode (ATM) or other packet-based transmission systems. The term "channel" as used herein should be understood to include any identifiable portion or portions of a communication medium which is used to transmit one or more signals and has an interference characteristic associated therewith,

and is thus intended to include, for example, a sub-channel, segment or other portion of a larger channel.

[0014] FIG. 1 illustrates channel classification UEP in accordance with an illustrative embodiment of the invention. In this embodiment, which is particularly well-suited for use in HIBOC DAB applications, the channels correspond generally to frequency bands, and the UEP technique is therefore referred to as frequency division UEP. Unlike certain of the approaches described in the above-cited EP-A-0936772, which can generally be characterized as time division UEP in which enhanced error protection may be provided for a certain class or classes of audio bits transmitted using a number of different channels, frequency division UEP in accordance with the invention provides enhanced error protection for a given class of bits by assigning that class of bits to a particular channel for transmission.

[0015] In the embodiment of FIG. 1, a portion of a frequency spectrum in an exemplary HIBOC DAB system is shown, including an analog host FM signal 100 with associated lower sidebands 102L, 104L and corresponding upper sidebands 102U, 104U. The sidebands represent portions of the frequency spectrum used to transmit digital audio information, and the sets of sidebands 102L, 102U and 104L, 104U correspond generally to frequency channels 102, 104, respectively, used to transmit the digital audio information. In accordance with the invention, a determination is made as to the interference characteristics associated with each of the frequency channels 102 and 104. This determination may be based, for example, on actual measurements of average signal-to-interference ratios within the channels, on known or estimated interference levels, or on any other information which provides an indication of relative or absolute interference levels for the channels. For example, it has been estimated based on previous experience with HIBOC systems that the portion of the spectrum of FIG. 1 at the highest and lowest frequencies is typically more susceptible to interference than the portion closest to the analog host FM signal 100. It will therefore be assumed that one of the channels, i.e., channel 102 in this example, has been determined to be less susceptible to interference than channel 104.

[0016] The illustrative embodiment of the invention, after determining the relative or absolute interference levels associated with n channels, where $n \geq 2$, to be used for transmission of digital audio information, separates the audio information into n classes of bits based on error sensitivity, and then assigns the n classes of bits to the n channels such that the bits most sensitive to errors are transmitted in the channels which are least susceptible to interference. In the FIG. 1 example, the audio information bits are separated into two classes, designated class I and class II, with class I including the bits most sensitive to errors. The determination of error sensitivity may be based on perceptual audio coding considerations such as those described in the above-cited EP-A-0936772. For example, class I may include the audio control bits as well as certain audio data bits corresponding to frequency bands which are perceptually important in reconstructing the encoded audio signal. These and other error sensitivity classification techniques are described in greater detail in EP-A-0936772 and will not be further described herein.

[0017] In the FIG. 1 example, the most sensitive audio information bits, i.e., class I, are transmitted in channel 102, i.e., the channel determined to be less susceptible to interference. This provides an increased robustness for the class I bits against the higher interference levels in channel 104. The two-class frequency division UEP approach illustrated in FIG. 1 will provide improvements over a conventional EEP approach. In one possible implementation of the FIG. 1 approach, the same channel code may be used for both the class I and II bits, but with a separate interleaving in time and frequency. It should be noted that the above-described frequency division UEP approach generally provides no improvement for channels which have a uniform interference level as a function of frequency. However, by taking into account the different interference characteristics of the channels, it can provide UEP for different classes of bits using the same code.

[0018] FIG. 2 illustrates another possible implementation of a two-class frequency division UEP approach in accordance with the invention. This example uses complementary punctured-pair convolutional (CPPC) codes, such as those described in greater detail in EP-A-0930738.

[0019] In this example, the bits in classes I and II are each separately coded using a rate-2/5 code which is formed as a combination of a pair of rate4/5 CPPC codes. These rate4/5 codes are referred to as half-bandwidth codes, and combine to form a rate-2/5 error correction code referred to as a full-bandwidth code. As is described in EP-A-0930738, a rate-1/3 mother code can be punctured to meet these exemplary HIBOC code requirements. The rate-1/3 mother code may be a rate-1/3 convolutional code having a constraint length K = 7 as described in J. Hagenauer, "Rate-compatible punctured convolutional codes (RCPC codes) and their applications," IEEE Transactions on Communications, Vol. 36, No. 7, pp. 389-400, April 1988.

[0020] The code rate is the ratio of input bits to output bits for the convolutional encoder, i.e., a rate-1/3 encoder generates three output bits for each input bit. A group of three coded output bits is referred to as a symbol. The value of K refers to the number of uncoded input bits which are processed to generate each output symbols. For example, a rate-1/3 convolutional encoder with K = 7 generally includes a seven-bit shift register and three modulo-two adders. The inputs of the each of the adders are connected to a different subset of the bits of the shift register. These connections are specified by the "generators" of the encoder. Because a given output symbol in this example is generated using the latest input bit as well as the previous six input bits stored in the shift register, the K = 7 encoder is said to have a "memory" of six. The rate-1/3, K = 7 code used in this example has the following three generators:

$$g_0 = 1011011$$

$$g_1 = 1111001$$

$$g_2 = 1100101$$

Each of the generators may be viewed as specifying the connections between bits of the seven-bit shift register and inputs of one of the modulo-2 adders. For example, the adder corresponding to generator g_0 generates the first bit of each output symbol as the modulo-2 sum of the bits in the first, third, fourth, sixth and seventh bit positions in the shift-register, with the first bit position containing the latest input bit. Similarly, the generators g_1 and g_2 generate the second and third bits, respectively, of each output symbol as modulo-2 sums of the bits in the positions designated by the respective generator values. The free Hamming distance d_f of the rate-1/3, K = 7 code with the above-noted generators is 14, and its information error weight c_{d_f}/P is one. When this code is punctured in a rate-compatible manner to rates of 4/11, 4/10, 4/9 and 1/2, the resulting rate-1/2 code is also the best rate-1/2, K = 7 convolutional code. Additional details regarding specific CPPC codes suitable for use in embodiments of the invention, as well as bit placement strategies for arranging the bits within the upper and lower sideband portions of the channels, can be found in EP-A-0930738.

[0021] FIGS. 3 and 4 illustrate other embodiments of the invention in which a dynamic boundary between class I and class II bits is used. In each of these embodiments, the boundary between class I and class II is as indicated by the dashed line 110. The portion of the frequency spectrum shown in FIGS. 3 and 4 includes the analog host FM signal 100, along with a lower sideband 106 and an upper sideband 108. As in the examples of FIGS. 1 and 2, the upper and lower sidebands are used to transmit digital audio information. In the FIG. 3 embodiment, the channels do not correspond directly to specific portions of the upper and lower sidebands. Instead, a first channel is defined by a portion of both the upper and lower sideband to one side of the dashed line 110, while a second channel is defined by the portion of the upper and lower sideband to the other side of the dashed line 110. Each of the upper and lower sidebands 106 and 108 uses, e.g, the same rate-2/5 code, as indicated. The use of a dynamic boundary allows a channel occupying a greater portion of the available frequency spectrum to be used to transmit class I bits. FIG. 4 shows another possible implementation using a dynamic boundary 110. A control channel or other suitable mechanism may be used to inform the receiver in a particular geographical area which configuration, e.g., the configuration of FIG. 3, the configuration of FIG. 4, or another type of configuration, is being used at the transmitter. The configuration may vary as a function of factors such as time or position within a coverage area.

[0022] It should be noted that in the embodiments of FIGS. 1 through 4, the same code, e.g., the same CPPC code pair, may be used for both classes I and II, or different codes may be used for each of the classes. In addition, as previously noted, the techniques can be readily extended in a straightforward manner to n channels and classes, where $n \geq 2$. Other possible variations include, for example, separate or joint interleaving, soft combining or equal gain combining, fixed or variable bit assignments, and use of other types of codes such as block codes.

[0023] FIG. 5 is a block diagram of an exemplary communication system 200 which implements the above-described frequency division UEP in accordance with the invention. The system 200 includes a transmitter 202 and a receiver 204 which communicate over an n-channel transmission medium 206. The transmitter 202 includes an audio encoder 210, e.g., a PAC encoder, for generating a sequence of audio packets from an analog audio input signal. Although this embodiment uses audio packets, such as those generated by a PAC encoder, the invention is more generally applicable to digital audio information in any form and generated by any type of audio compression technique. The audio packets from encoder 210 are applied to a classifier 212 which converts the packets into separate bit streams corresponding to n different classes of audio information bits. The classifier 212 is also responsible in this embodiment for assigning each of the classes of bits to one of the available channels such that the classes of bits most sensitive to errors are transmitted in the channels which are least susceptible to interference, as previously described. The separate bit streams from the classifier 212 are applied to a set of channel coders 214. The symbol outputs of the channel coders 214 are supplied to a set of interleavers 215 which provide interleaving of the symbols within each channel over multiple audio packets. The interleaved symbols are then supplied to a set of orthogonal frequency division multiplexed (OFDM) modulators 216 for modulation in accordance with conventional OFDM techniques. The OFDM modulators may provide, for example, single-carrier modulation in each of the channels. Of course, other types of modulation may be used in alternative embodiments.

[0024] The transmitter 202 may include additional processing elements, such as a multiplexer, an upconverter and the like, which are not shown in FIG. 5 for simplicity of illustration. In addition, the arrangement of elements may be varied in alternative embodiments. For example, other types of modulators may be used in place of the OFDM mod-

ulators 216, such as modulators suitable for generating signals for transmission over a telephone line or other network connection, and separate interleaving and coding need not be applied to each of the channels.

[0025] The receiver 204 receives the transmitted OFDM signals from the transmission channels 206, and processes them in OFDM demodulators 219 to recover the interleaved symbols for each of the channels. The symbols are deinterleaved in a set of deinterleavers 220, and then applied to a set of channel decoders 222. The bit streams at the output of each of the decoders in the set of decoders 222 correspond to the different classes of audio information bits. These bit streams are then processed in a declassifier 224 which reconstructs audio packets from the bit streams. The resulting sequence of audio packets are then decoded in an audio decoder 226 to reconstruct the original analog audio signal.

[0026] Like the transmitter 202, the receiver 204 may include additional processing elements which are not shown in FIG. 5. It should also be noted that various elements of the system 200, such as the interleavers 215 and the deinterleavers 220, may be eliminated in alternative embodiments. Moreover, various elements of the system 200, such as the audio encoder 210 and decoder 226, the channel coders 214 and decoders 222, and the classifier 212 and declassifier 224, may be implemented using an application-specific integrated circuit, microprocessor or any other type of digital data processor, as well as portions or combinations of such devices. Various aspects of the invention may also be implemented in the form of one or more software programs executed by a central processing unit (CPU) or the like in the- digital data processor.

[0027] Simulation results for an exemplary frequency division UEP (FD-UEP) system such as that described in conjunction with FIGS. 1-5 are shown in TABLE 1 below. In the simulations, a channel was assumed to include two disjoint segments, designated segment I and segment II. Such segments are also referred to herein as sub-channels, and it should be noted that each segment is itself considered to fall within the general definition of "channel" given above. In other words, each segment may be considered a channel. With a suitable interleaver depth, the channel quality may be assumed to be constant over a particular segment. The two segments can thus be parameterized by an interference characteristic such as, for example, the corresponding signal-to-noise level measured in terms of E_S/N_0. Gaussian channel conditions are assumed in the simulations.

[0028] In an EEP transmission system operating over segments I and II, it is reasonable to assume half of the channel coded bits encounter a channel condition which exists in segment I and another half encounter conditions existing in segment II. For the FD-UEP system, it is assumed that audio information bits are separated into a class I which includes control bits and a first portion of the audio data bits, and a class II which includes a second portion of the audio data bits. These classes I and II may correspond, for example, to classes 1* and 2*, respectively, as described in EP-A-0936772. In accordance with the present invention, the class I and II bits may be interleaved and transmitted independently over segments I and II, respectively. Therefore, class I bits are exposed to the channel condition in segment I and class II bits face the channel condition in segment II. In each of the simulations, a convolutional channel code with a rate of 2/5 was used, as described above, and the same outer cyclic redundancy codes (CRCs) were also used.

TABLE 1

Simulation No.	Channel Condition (E_S/N_0) in dB		EEP Quality	FD-UEP Quality
	Segment I	Segment II		
1.	-0.5	-0.5	Slight distortion	Slight distortion
2.	-0.5	-2.5	Partial Breakdown (~ 50% Muting)	Some distortions Audio BW reduction Some noise bursts
3.	-0.5	-3.0	Total Breakdown (> 75% Muting)	Some distortions Audio BW reduction

[0029] Subjective audio quality for the above-described EEP and FD-UEP systems were evaluated for different channel conditions, and the qualitative results are summarized in TABLE 1. As expected, if the channel conditions on the two segments are roughly equivalent, as in simulation 1 in TABLE 1, both EEP and FD-UEP systems perform similarly. On the other hand, it is clear from simulations 2 and 3 in TABLE 1 that when the conditions in the two segments are substantially different, the FD-UEP system exhibits a much more graceful degradation. More specifically, if a given channel condition exists in segment I and segment II is approximately 2.0 dB worse, the EEP system is unacceptable with muting nearly half the time. The FD-UEP system, in contrast, survives with only reduced audio bandwidth and some increase in distortions. When the channel condition in segment II is about 2.5 dB worse than that in segment I, the EEP system mutes more than 75% of the time, while the FD-UEP system survives albeit with lower audio bandwidth and increased distortions. In other words, as the interference level in segment II increases, the audio quality in the FD-

UEP system "bottoms out" at a lower yet often acceptable quality level. By way of comparison, the EEP system mutes almost completely under these same conditions.

[0030] The distortions noticed in the FD-UEP system in simulations 2 and 3 of TABLE 1 are primarily due to audio bandwidth reduction and aliasing attributable to the classifier described in EP-A-0936772. If the difference in the channel conditions between segment I and II is relatively moderate, there is one other potential distortion as noticed in simulation 2, i.e., an occasional burst of high frequency noise. This happens when channel conditions in segment I are much beyond the point of failure for class II bits, i.e., >20% PAC packet loss for these bits, yet not severe enough, i.e., <50-60% PAC packet loss, to lead to a complete muting for class II in the PAC error mitigation algorithm. This may lead to a situation in which the performance of the FD-UEP system may actually improve slightly when the channel condition in segment II becomes progressively worse beyond a certain threshold. It should be noted that in spite of the above-described distortions, the simulations clearly indicate that an FD-UEP system in accordance with the invention is preferable to an EEP system at least in terms of providing a more graceful performance degradation.

[0031] The above-described embodiments of the invention are intended to be illustrative only. For example, the invention can be applied to the transmission of digital information other than audio, such as video, images and other types of information. In addition, alternative embodiments of the invention may utilize different types of channels. Different types of coding, e.g., convolutional coding with different memories or other characteristics, or other types of codes such as block codes, may also be used. Furthermore, the invention may make use of different types of modulation, including, e.g., single-carrier modulation in every channel, or multi-carrier modulation, e.g., OFDM, in every channel. A given carrier can be modulated using any desired type of modulation technique, including, e.g., a technique such as m-QAM, m-PSK or trellis coded modulation.

[0032] It should be noted that any of the error sensitivity classification techniques described in the above-cited EP-A-0936772, including multipacket error protection profiles, may be used to classify the information bits in terms of error sensitivity. The UEP techniques described in EP-A-0936772 may be used to provide further levels of UEP within a given class, e.g., within a class assigned to a channel having a substantially uniform interference level. In addition, the techniques of the invention may be used to provide any number of different classes of UEP for information, and may be used with a wide variety of different bit rates and transmission channels. For example, as previously noted, alternative embodiments can extend the illustrative two-class techniques described above to any desired number n of classes in a straightforward manner.

[0033] Further embodiments of the invention could use other techniques for providing adaptive numbers and types of different classes and channels. In addition, the number and/or characteristics of the channels and classes, as well as the assignment of classes to channels, may be fixed or dynamic. For example, if the interference characteristics associated with the channels vary as a function of time or position within a coverage area, the assignment of the classes of bits to the channels can be varied as a function of time so as to ensure that the classes of bits having the greatest error sensitivity continue to be transmitted over the channels which are least susceptible to interference as the interference characteristics vary. As another example, the bandwidth or other characteristic of a particular channel or channels may be made to vary as a function of time. These and numerous other alternative embodiments and implementations within the scope of the following claims will be apparent to those skilled in the art.

Claims

1. A method of processing information bits for transmission in a communication system, the method comprising the steps of:

    separating the information bits into n classes of bits based on error sensitivity, where n is greater than or equal to two; and
    assigning the classes of bits to n channels, each having an interference characteristic associated therewith, such that the class of bits having the greatest error sensitivity is transmitted over the channel which is the least susceptible to interference.

2. The method of claim 1 further including the step of determining the interference characteristics associated with the n channels to be used in transmitting the information bits.

3. An apparatus for use in processing information bits for transmission in a communication system, the apparatus comprising:
    a transmitter operative to separate the information bits into n classes of bits based on error sensitivity, where n is greater than or equal to two, and to assign the classes of bits to n channels having corresponding interference characteristics, such that the class of bits having the greatest error sensitivity is transmitted over the channel which

is the least susceptible to interference.

4. The method of claim 1 or apparatus of claim 3 wherein the interference characteristics are determined by measuring interference levels for at least a subset of the channels.

5. The method of claim 1 or apparatus of claim 3 wherein the interference characteristics are determined by obtaining information regarding known interference levels for at least a subset of the channels.

6. The method of claim 1 or apparatus of claim 3 wherein each of the channels corresponds to one of a different frequency band, a different time slot, and a different code division slot.

7. The method of claim 1 or apparatus of claim 3 wherein the interference characteristics associated with at least a subset of the channels vary as a function of time, and the assignment of the classes of bits to the channels is varied so as to ensure that the classes of bits having the greatest error sensitivity continue to be transmitted over the channels which are least susceptible to interference as the interference characteristics vary.

8. The method of claim 1 or apparatus of claim 3 wherein the interference characteristics associated with at least a subset of the channels vary as a function of position within a coverage area, and wherein the assignment of the classes of bits to the channels is varied so as to ensure that the classes of bits having the greatest error sensitivity continue to be transmitted over the channels which are least susceptible to interference as the interference characteristics vary.

9. The method of claim 1 or apparatus of claim 3 wherein an amount of channel resources used for a particular class of information bits varies as a function of time.

10. The method of claim 1 or apparatus of claim 3 wherein at least one of the channels has a substantially uniform interference level, and unequal error protection is provided for the class of information bits within the at least one channel.

11. A method of processing information bits for transmission in a communication system, the method comprising the steps of:

    separating the information bits into n classes of bits based on error sensitivity, where n is greater than or equal to two; and
    selecting a given one of n channels for transmitting a corresponding one of the n classes of bits, based on relative interference characteristics of the n channels.

12. An apparatus for use in processing information bits for transmission in a communication system, the apparatus comprising:
    a transmitter operative to separate the information bits into n classes of bits based on error sensitivity, where n is greater than or equal to two, and to select a given one of n channels for transmitting a corresponding one of the n classes of bits, based on relative intcrference characteristics of the n channels.

FIG. 1

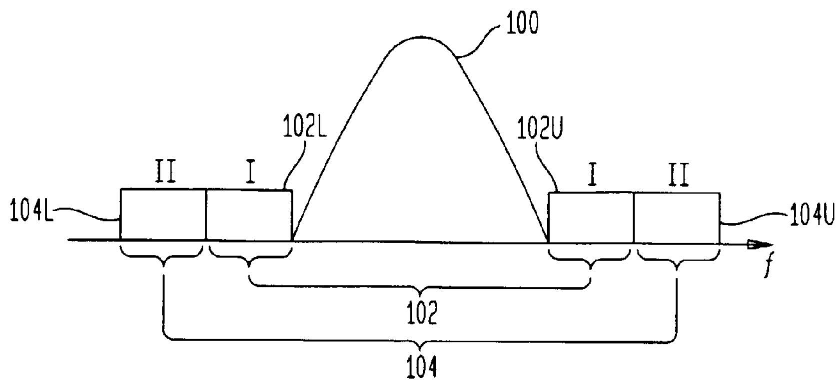


FIG. 2

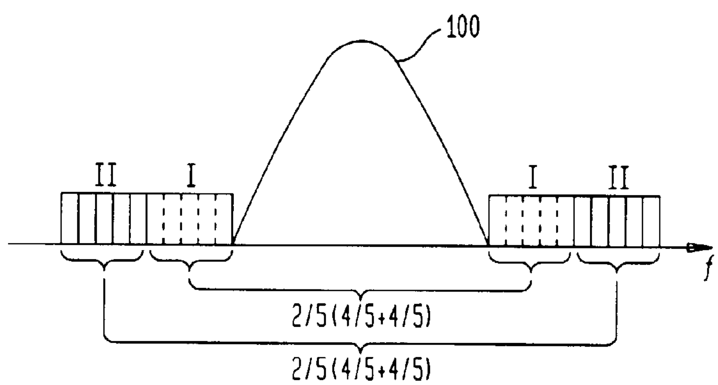


FIG. 3

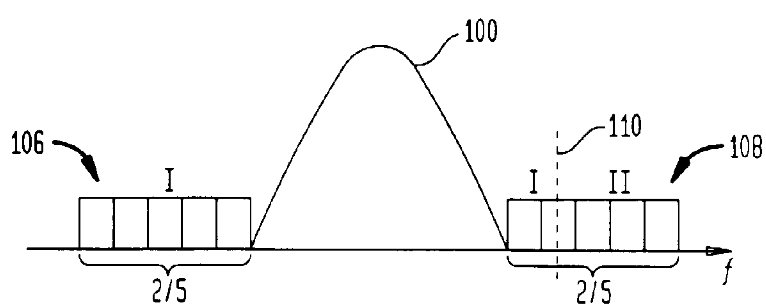


FIG. 4

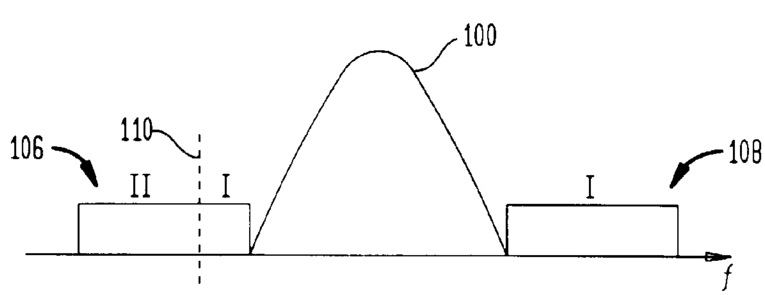
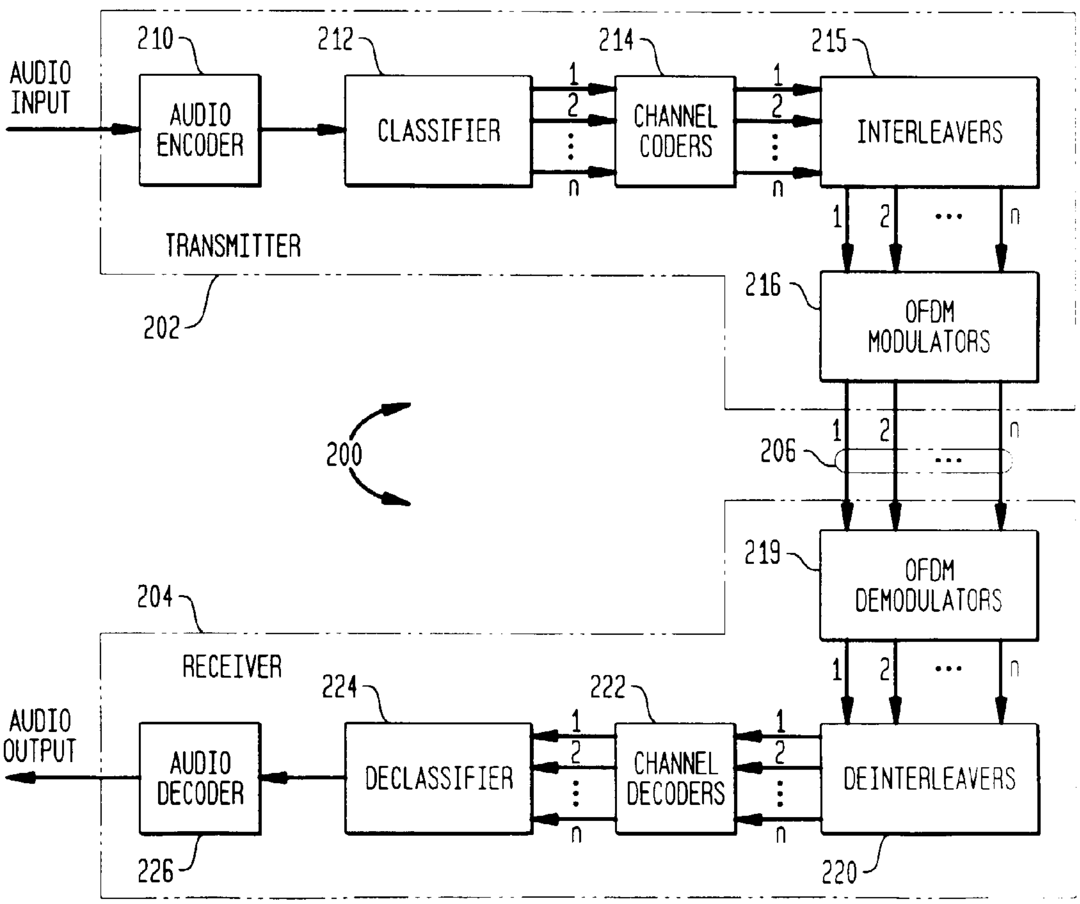


FIG. 5



AUDIO INPUT
210
AUDIO ENCODER
212
CLASSIFIER
214
CHANNEL CODERS
215
INTERLEAVERS
1
2
n
TRANSMITTER
202
216
OFDM MODULATORS
200
206
219
OFDM DEMODULATORS
204
RECEIVER
224
222
AUDIO OUTPUT
AUDIO DECODER
DECLASSIFIER
CHANNEL DECODERS
DEINTERLEAVERS
226
220

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(71) Applicant:
Sony International (Europe) GmbH
50829 Köln (DE)

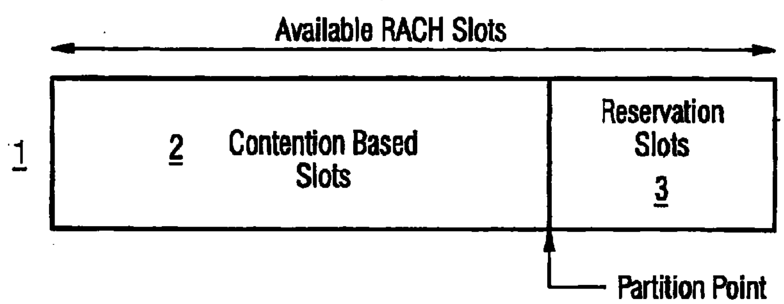
(72) Inventors:
- **Halton, John**
c/o Sony International (Europe) GmbH
70736 Fellbach (DE)
- **James, Paul**
c/o Sony International (Europe) GmbH
70736 Fellbach (DE)

(74) Representative:
Körber, Martin, Dipl.-Phys. et al
Mitscherlich & Partner
Patentanwälte
Sonnenstrasse 33
80331 München (DE)

(54) **Random access channel partitioning scheme for CDMA system**

(57) The present invention relates to a method and to devices for transmitting and receiving data in a code division multiple access telecommunication system. A random access time window is provided, which comprises a plurality of random access slots for transmitting random access data for example from mobile station to a base station. The random access time window is partitioned in a first section (2) and a second section (3), whereby the first section contains contention based random access slots and the second section contains reservation based random access slots. Thereby, the random access channel can be adopted to and used in a flexible manner in dynamically changing environments.

FIG 2



EP 0 993 211 A1

Description

[0001] The present invention relates to a method and to devices for transmitting and receiving data in a code division multiple access telecommunication system.

[0002] A telecommunication system is a system, in which data are communicated between one or more base stations and one or more mobile stations. Thereby, the communication area is divided in cells, in which one base station communicates with one or more mobile stations. Multiple access systems are used to support the simultaneous access of a plurality of mobile stations to one base station within the limited resources of the transmission system. Several multiple access systems are known, e. g. frequency division multiple access (FDMA), time division multiple access (TDMA), or code division multiple access (CDMA). Additional to these basic types of multiple access systems, combinations of these different systems are possible and in practical use. The GSM-system for example uses a combination of FDMA and TDMA.

[0003] The present invention particularly relates to the transmission and reception of random access data in a code division multiple access system. Random access data are transmitted in the so-called random access slot (RACH) from a mobile station to build up a connection. The random access data from the mobile station contain a request, if the base station has sufficient resources available to build up the required connection or to transfer user data.

[0004] The random access slot comprises or consists of succeedingly or periodically provided random access time windows, in which several random access slots are available. The different random access slots are randomly chosen by a mobile station for the transmission of random access data. In a currently proposed wide band direct sequence CDMA (WCDMA) system, the random access time windows are based upon an initial preamble scrambling code, which differentiates between one cell and another cell. Thereby, these codes need to be planned to ensure that neighboring cells do not use the same code. Therefore, within the preamble part of each random access slot burst, is provided the preamble signature, which is one of 16 separate codes available for use within that cell. These 16 codes can be seen as separate slots. One of these codes is chosen randomly by the mobile station for the transmission of random access data. Beforehand, the base station broadcasts, which codes are available in each cell over the broadcast control channel (BCCH). In addition, within one time frame (10ms) are provided 8 time offsets, each of 1,25 ms, allowing a further 8 variations. In other words, in each time frame a random access time window is provided, which comprises a plurality of random access slots for transmitting random access data from one or more mobile stations to the base station. The random access time window thereby extends over the time frame of 10ms, so that 128 different random access slots (16 separate preamble codes and 8 time offsets) are provided within one random access time window.

[0005] A collision, i. e. a situation, where messages from two or more mobile stations collide and are lost, only occurs, when both the preamble code and the time offset are chosen in the same random access time window. In practice, it is expected that only about 30% of the theoretical maximum of 128 accesses per 10ms will be possible.

[0006] In a situation, where a number of packet data users are operating in a bursty traffic situation, this maximum could be quickly reached. In such a situation the access to the network will either become slower or not be possible at all. This is due to congestion caused by the build up of first time requests and the retransmissions made necessary by previous collisions. Since the access to the random access slots is only contention based as shown in Fig. 1, a guaranteed upper limit on the amount of time needed to access the system even after an initial burst is not ensured. For packet data applications, which demand a constant periodic delivery of data, ready access is critical.

[0007] WO 98/24250 discloses a TDMA system, which allocates a fixed random access slot constantly to a particular mobile station.

[0008] The object of the present invention is thus to provide a method and devices for transmitting and receiving data in a code division multiple access telecommunication system, in which a random access time window comprising a plurality of random access slots for transmitting random access data is provided and which enable a reduced congestion on the random access slots and a more efficient use of the random access time window. The importance of an effective random access slot utilization arises particularly from increased load from packet data capacity requests and the transfer of small amounts of user data within random access slot bursts.

[0009] This object is achieved by a method for transmitting and receiving data in a code division multiple access telecommunication system, comprising the steps of providing a random access time window comprising a plurality of random access slots for transmitting random access data from at least one first communication device to a second communication device, partitioning the random access time window in a first and a second section, whereby the first section contains contention based random access slots and the second section contains reservation based random access slots. The first communication device can thereby be a mobile station and the second communication device can thereby be a base station of the telecommunication system. Reservation based random access slots are slots which cannot be accessed on a contention basis, but only if they had been reserved before.

[0010] This object is further achieved by a device

for transmitting and receiving data in a code division multiple access telecommunication system, in which a random access time window comprising a plurality of random access slots for transmitting random access data is provided, the random access time window being partitioned in a first and a second section, whereby the first section contains contention based random access slots and the second section contains reservation based random access slots, with means for randomly choosing a random access slot from said first section and means for transmitting random access data in said chosen random access slot. This device for transmitting and receiving data is e. g. a mobile station of the telecommunication system.

[0011] The above object is further achieved by a device for transmitting and receiving data in a code division multiple access telecommunication system, in which a random access time window comprising a plurality of random access slots for transmitting random access data is provided, with means for partitioning the random access time window in a first and a second section, whereby the first section contains contention based random access slots and the second section contains reservation based random access slots, and means for transmitting partitioning information defining the partitioning of said random access time window. This device for transmitting and receiving data can be e. g. a base station of the multiple access telecommunication system.

[0012] The partitioning of the random access time window into the first section containing contention based random access slots and a section containing reservation based random access slots thereby allows to reduce congestion in the random access slots and to more efficiently utilize the resources of the random access time window. Advantageously, the number of random access slots in the first and the second section, respectively, can be set variably depending on system requirements. This allows to flexibly modify the random access time window from one moment to the next moment to adopt the random access time window to dynamically changing environments and requirements. Each random access slot in said random access time window can be defined by a time offset value and a preamble code. Further advantageously, said second communication device (which e. g. can be a base station) periodically transmits partitioning information defining the partitioning of the random access time window to at least one first communication device, which for example can be a mobile station. Thereby, the partitioning of the random access time window can be flexibly adopted to system requirements. The first communication device, which for example can be a mobile station, randomly chooses one of said random access slots from said first section of said random access time window for transmitting random access data to the second communication device, which for example can be a base station.

[0013] Advantageously, said random access data transmitted in a random access slot of said first section comprises reservation data for reserving at least one random access slot of said second section in at least one succeeding random access time window. Thereby, ready access for packet data applications is ensured, which demand a constant delivery of data. In this case, said reservation data can comprise data on the time duration required for the total number of reserved random access slots to indicate the end of the reservation. Thereby, the reservation of random access slots in the second section of succeeding random access time windows ends automatically upon expiring of the pre-set time period. Said reservation data can further comprise information on a number of random access slots to be reserved in the first succeeding random access time window and information on a continuos reduction of said number in the following random access time windows. In this case, the number of reserved random access slots in the second sections of succeeding random access time windows can be decreased gradually until the end of the reservation term.

[0014] Said first communication device, which can for example be a mobile station, can transmit reservation termination data in a first or second section of a random access time window for indicating the end of the reservation of the required random access slot. This might be necessary in cases, in which already reserved random access slots in the second sections of succeeding random access time windows are not necessary and have to be released for usage by other users. Alternatively, said second communication device, which can for example be a mobile station, terminates the reservation upon determination of a non-usage of reserved random access slots. The second communication device can thereby automatically terminate the reservation if it is determined, that reserved random access slots are not used. In this case, the second communication device can transmit information on the termination of the reservation to the corresponding first communication device to inform that earlier reserved random access slots in second sections of succeeding random access time windows are not reserved anymore.

[0015] The device for transmitting and receiving data according to the present invention, which might be a mobile station and comprises means for randomly choosing a random access slot from said first section and means for transmitting random access data in said chosen random access slot can further comprise means for receiving periodically transmitted partitioning information defining the partitioning of said random access time window.

[0016] The device for transmitting and receiving data according to the present invention, which might be a base station, and comprises means for partitioning the random access time window in a first and a second section and means for transmitting partitioning information defining the partitioning of said random access time window, might further comprise means for receiving

random access data in one of said random access slots from said first section of said random access time window. Said means for partitioning the random access time window advantageously sets the number of random access slots in said first and second section, respectively, variably depending on system requirements, whereby said means for transmitting partitioning information periodically transmit said partitioning information. Said device for transmitting and receiving data, which might be a base station, can further comprise means for determining a non-usage of reserved random access slots, whereby the reservation is terminated upon a positive result of said determination. Said means for transmitting information can thereby transmit information on the termination of the reservation upon a positive result of said determination.

[0017] In the method and devices for transmitting and receiving data according to the present invention, the random access time window is partitioned into a first and a second section, whereby the first section contains contention based random access slots, and the second section contains reservation based random access slots. The reservation based random access slots can be dedicated to a communication device, e. g. a mobile station, for a limited time. A request for guaranteed reservation of reservation based random access slots is made through an initial contention based random access slot. Thereby, the network or the base station may also assign a reserved random access slot. These reserved random access slots are then used for uplink data transfer between a mobile station and a base station or requests for a channel to transmit user data. Since the partitioning of the random access time window can be set variable depending on system requirements, the random access slot resources in the random access time window can be used more efficiently, e. g. by minimizing contention based access, since bandwidth is not used for retransmissions when bursts collide. Furthermore, a breakdown can be avoided, which can occur, when the number of retries and subsequent retransmissions increase the number of collisions until deadlock occurs. The telecommunication network or the base station can dynamically control the partitioning of the random access resources in the random access time window into contention and reservation based channels. Thus, the exact partition can be tailored to the particular requirements from moment to moment. This partitioning information is broadcasted to the mobile stations upon the broadcast control channel (BCCH). The present invention thereby particularly enables a faster access to the communication network particular in cases where a number of packet data users are operating in bursty traffic situations. Further, congestion caused by the build up of first time requests and retransmissions made necessary by previous collisions can be avoided.

[0018] In the following, preferred embodiments of the present invention are explained in detail referring to the enclosed figures, in which

Fig. 1 shows a schematic diagram of a known random access time window,

Fig. 2 shows a schematic diagram of a first example of a random access time window according to the present invention,

Fig. 3 shows a schematic diagram of a second example of a random access time window according to the present invention,

Fig. 4 shows a schematic diagram of a third example of a random access time window according to the present invention,

Fig. 5 shows a schematic diagram of a random access time window succeeding the random access time window shown in Fig. 4,

Fig. 6 shows a schematic diagram of a fourth example of a random access time window according to the present invention,

Fig. 7 shows a diagram of a reservation procedure for reserving reservation based random access slots according to the present invention,

Fig. 8 shows four schematic diagrams of random access time windows according to the present invention for explaining the use of contention based random access slots and reservation based random access slots according to the present invention,

Fig. 9 shows a schematic diagram of a plurality of succeeding second sections of random access time windows according to the present invention for explaining the duration based expiring of reserved random access slots,

Fig. 10 shows a schematic diagram of a plurality of succeeding second sections of random access time windows according to the present invention for explaining the mobile station originated cancellation of reserved random access slots,

Fig. 11 shows a schematic diagram of a plurality of second sections of random access time windows according to the present invention for explaining the network originated cancellation of reserved random access slots,

Fig. 12 shows a schematic diagram of a plurality of succeeding second sections of random access time windows according to the present invention for explaining the decay based expiring of reserved random access slots,

Fig. 13 shows a block diagram of a base station and two mobile stations incorporating the present invention,

Fig. 14 shows a block diagram of a mobile station incorporating the present invention, and

Fig. 15 shows a block diagram of a base station incorporating the present invention.

[0019] In Fig. 1, a schematic diagram of a known random access time window is shown. In this known

random access time window, all available random access slots are contention based. The number of the available contention based random access slots is defined by 8 different time offset values and 16 different preamble codes, so that a total of 128 different random access slots is available per random access time window. In the present application, a random access time window corresponds to the duration of a random access time frame, e. g. 10ms. It is, however, to be noted, that a random access time window according to the present invention can be defined by other parameters as long as a plurality of random access slots for transmitting random access data are available within a certain time period. Further, throughout the following description, the present invention is explained relating to a telecommunication system comprising a base station as for example shown in Fig. 15 and one or more mobile stations as for example shown in Fig. 14.

[0020] In Fig. 2, a schematic diagram of a first example of a random access time window 1 according to the present invention is shown. The shown random access time window 1 is partitioned into a first section 2 and a second section 3. The first section 2 contains contention based random access slots and the second section 3 contains reservation based random access slots. The contention based random access slots are used for initial access, singular packet transfer and for reservation of additional dedicated random access slots for a time period within the second section of succeeding random access time windows. The reservation based random access slots in the second section of the random access time window 1 can be allocated to mobile stations, which periodically on a regular basis wish to transfer uplink packet data or make a request for channel reservations to transfer data.

[0021] The partition point shown in the random access time window 1 of Fig. 2 allocates a larger number of random access slots to the first section 2 than to the second section 3. Hereby, the partition point only divides the random access time window 1 in view of the preamble codes. In other words, a certain part of the preamble codes available, e. g. 10 of a total of 16 preamble codes, is allocated to the first section 2, whereas the rest of the preamble codes, e. g. the remaining 8 of a total of 16 preamble codes, is allocated to the second section 3. In the example shown in Fig. 2, all time offset values, e. g. 8 different time offset values, remain available for the random access slots in the first section 2 as well as the second section 3.

[0022] The actual position of the partition point can be dynamically controlled by a telecommunication network depending upon system requirements, but with respect to current random access slot reservation commitments. In case that a large number of mobile stations attempts to access the network, the number of contention based random access slots in the first section 2 is increased by moving the partition point to reduce the number of reservation based random access slots in the second section 3. If the demand for packet based data is high, the partition point may be moved to reduce the number of contention based random access slots in the first section 2 and create more available reservation based random access slots in the second section 3. This possibility is shown as a second example in the schematic diagram of a random access time window in Fig. 3. In the example shown in Fig. 3, the first section 2 contains a smaller number of contention based random access slots than the second section 3 contains reservation based random access slots.

[0023] In Fig. 4, a third example of a random access time window according to the present invention is shown. In the example shown in Fig. 4, the partitioning point divides the first section 2 from the second section 3 by partitioning the time offset values as well as the preamble codes, so that the random access time window 1 is partitioned in two dimensions. In the example shown in Fig. 4, only 6 of a total of 8 time offset values and 8 of a total of 16 preamble codes are available for the reservation based random access slots in the second section 3.

[0024] In Fig. 5, a random access time window is schematically shown, which is partitioned identically to and succeeds the random access time window shown in Fig. 4. One of the random access slots of the second section 3 of the time window shown in Fig. 5, which is identified by the reference numeral 16, is a reserved random access slot, which has been reserved by one of the contention based random access slots of the first section 2 of the time window shown in Fig. 4.

[0025] In Fig. 6, a fourth example of a random access time window 1 according to the present invention is shown, which illustrates another partitioning of the random access time window. The random access time window shown in Fig. 6 is divided in two dimensions, whereby 8 of a total of 16 preamble codes and 4 of a total of 6 time offset values are allocated to the second section 3.

[0026] In Fig. 7, the signaling procedure for the reservation of reservation based random access slots is explained. A mobile station 27 as for example shown in Fig. 14 wishes to transmit packet-oriented data and therefore makes a request for a reservation based random access slot using a contention based random access slot. Therefore, the mobile station 27 randomly chooses a time offset value and a preamble code from the available contention based random access slots of section 1 of the current random access time window and transmits the required reservation information to the corresponding base station as for example shown in Fig. 15. The required reservation information can e. g. comprise the number of random access slots to be reserved, the start time of the required random access slots to be reserved and the total duration of the reservation. The base station 26 determines, if it has enough resources in the corresponding second section 3 of the succeeding random access time windows to grant the

reservation request from the mobile station 27. In other words, the base station 26 decides, if sufficient reservation based relevant access channels in the succeeding random access time windows are available and sends, if the decision is positive, a corresponding response, e. g. on the forward access channel, back to the mobile station 27. The response burst contains the reservation parameters, e. g. the time the reservation is available and the duration of the reservation. Should the base station 26 wish to allocate less random access slots in the second section 3 of the succeeding random access time windows as requested by the mobile station 27 or not be able to grant the request at all, this can also be indicated within the response burst.

[0027] In case that the response burst from the base station 26 grants all the required reservation resources to the mobile station, the mobile station transmits e. g. packet orientated random access bursts using the preamble code and the time offset value provided in the burst response from the base station. For example, random access slot 16 shown in the second section 3 of the random access time window of Fig. 5 is a random access slot, which had been reserved by a corresponding request sent within a random access slot of the first section 2 of the preceding random access time window shown in Fig. 4.

[0028] In case that the base station determines or receives information that the number of the contention based random access slots need to be increased, it broadcasts the new partition point to the corresponding mobile stations, so that the succeeding random access time windows are partitioned correspondingly.

[0029] In Fig. 8, four schematic diagrams (a) - (d) are shown to explain how contention based random access slots of the first section 2 of a random access time window are used to reserve reservation based random access slots in the second section 3 of succeeding random access time windows. Although the diagrams of Fig. 8 are shown for a single time domain, it is to be noted, that the reservation principle takes place in the continuous time domain.

[0030] The mobile station 27 selects a preamble code and a time offset from the set of available contention based random access signals and constructs a random access burst. The random access burst also contains the required resource reservation, e. g. the required number of random access slots to be reserved, the required duration of the reservation, the desired reservation policy, and the desired release procedure. The random access slot burst may further contain requests for a number of independent reservations. The random access burst is then transmitted from the mobile station over the air interface to a base station 26. The diagram shown in Fig. 8(a) shows three simultaneously transmitted initial random access bursts 4, 5, 6, transmitted in the corresponding random access slots of the first section 2 of the current random access time window. The random access bursts 4, 5, 6 are received at the receiver controlling radio resources, e. g. a base station. Thereby, the available set of contention based random access slots and reservation based random access slots is presented on a broadcast channel by transmitting only the corresponding partition point from the base station to the mobile stations 27, 28. Thus, the partition point can be continuously modified and transmitted with a very low overhead.

[0031] The base station allocates, if available, the requested resources from the reservation based random access slots in the second sections 3 of the succeeding random access time windows and marks these in use beginning at the requested time and terminating after the requested duration. This is shown in the diagram of Fig. 8(b). For the ease of explanation, the initial random access burst 4, 5, 6 of the preceding random access time window are also shown. The random access burst 4 of the preceding random access time window reserves three random access slots 7, 8, 9 in the section 2 of the succeeding random access time window. The initial random access burst 5 reserves one random access slot in the second section 3 of the preceding random access time window. The initial random access burst 6 reserves one random access slot 10 in the second section 3 of the succeeding random access time window.

[0032] The actual succeeding random access time window 1 succeeding the random access time window 1 shown in Fig. 8(a) is shown in the diagram of Fig. 8(b). In this succeeding random access time window 1, no initial random access bursts are transmitted in the random access slots of the first section 2, but the reserved random access slots 7, 8, 9, 10, and 11 in the second section 3 of the random access time window are available for transmitting data. If the base station determines, that fewer resources as requested are available and the mobile station 27 indicated this as reasonable in the indicated desired reservation policy, a maximum of resources within the available resources is selected. If the desired reservation procedure indicated that an exact match must be made, a negative result is generated and no random access slots are reserved in the succeeding random access time windows. A notification of the requested reservation outcome is transmitted to the mobile stations. This includes the allocated random access slots in the second section 3 of the succeeding random access time windows and the valid duration for their use. On reception of the positive notification of a successful reservation, the mobile stations proceed to make random access bursts on the reserved random access slots. The result might be as shown in Fig. 8(c). For instance, a web browsing service may use the reserved random access slots to request specific web pages by transmitting the HTML address in the random access burst. Alternatively, the reserved random access slots may be used to request additional channels for up- or downlink data transfer. The diagram in Fig. 8(d) shows how random access slots, that have expired, may

be re-allocated to new requests. For example, a new initial random access burst 12 transmitted in the first section 3 of a random access time window may reserve a random access slot 14 with the same address as the old reserved random access slot 11 of Figs. 8(b) and 8(c), which has expired. Another new initial random access burst 13 may reserve a new random access slot 15 in the second section 3 of the same succeeding random access time window. Further on, as shown in Fig. 8(d), the random access slots 7 and 8, which had been reserved in an earlier random access time window, stay reserved also in the same random access time window as the reserved random access slots 14 and 15.

[0033] It is further possible, that the base station 26 automatically assigns a particular reservation based random access slot in each random access time window for a mobile station 27, which makes a large number of accesses to contention based random access slots. The mobile station is informed on the automatic assignment by transmission of the reserved random access slot by the base station.

[0034] Reserved random access slots must at some point be relinquished. It is important, that the scheme used should serve the needs of the resource allocator, the resource requester, and the characteristics of data being transmitted either through the random access burst or the establishment of dedicated transport channels. Therefore, relating to Fig. 9 - 12, a number of mechanisms are presented, which may be used when requesting and/or terminating reservation of random access slots. In Fig. 9 - 12, only reservation based random access slots of the second sections 3 of sequential random access time windows are shown.

[0035] Fig. 9 shows an example for a duration based reservation of random access slots in the second section 3. The perspective view in Fig. 9 thereby shows five sequential second sections 3 provided every 10ms. The initial reservation takes place in the first section 2 (not shown) of the random access time window, of which only the second section 3 is shown at 10ms. The initial reservation reserves a random access slot 17 in the three succeeding random access time windows shown at 20ms, 30ms, and 40ms. Since the required duration for the reservation based random access slots had been set within the original request, the mobile station is allowed to use the reserved random access slots up until the end of the duration. The end of the duration is reached at the 5^{th} random access time window, of which only the second section 3 is shown at 50ms. At this point, the reserved random access slot 17 is expired and a reservation must be repeated for further access using reservation based random access slots.

[0036] Fig. 10 shows an example for a mobile station originated cancellation of random access slots. The perspective view in Fig. 10 thereby shows three sequential second sections 3 provided every 10ms. The mobile station might wish to release its use of certain reserved slots immediately. In the example shown in Fig. 10, the succeeding slots 19 had been reserved and used by the mobile station to transmit packet oriented data. Although the corresponding succeeding random access slot 20 in the succeeding random access time window had been reserved initially, the mobile station wishes to cancel this reservation. This case may occur, when a mobile station speculatively asks for a number of future reservations, which subsequently are not required and are released prematurely. It is important, that reservation slots can be removed selectively. The mobile station makes a random access burst using the reservation code, which it wishes to cancel. On reception of the random access burst by the base station, the allocation is removed and the initially reserved random access slot is available to be reallocated.

[0037] Fig. 11 shows an example for a mobile station originated cancellation of random access slots. The perspective view in Fig. 11 thereby shows five sequential second sections 3 provided every 10ms. A mobile station has reserved succeeding random access slots 21, 22, and 23 of succeeding random access time windows. Of these reserved random access slots, the mobile station uses only the random access slots 21 in the first two succeeding random access time windows to transmit data. The succeeding two random access slots 22 are not used by the mobile station, which is detected by the base station. In other words, if a burst on reserved random access slots has not been received during a predetermined number of time periods, the base station cancels the succeeding reserved random access slots. In the case shown in Fig. 11, the base station cancels the reservation for the succeeding reserved random access slots 23 in the fifth random access time window, since the two preceding random access slots 22 have not been used by the mobile station. It may be necessary for the base station to confirm the cancellation by transmission of a cancellation indication to the mobile station to avoid later collisions within the reserved random access slots. This may occur, when a mobile station later attempts to use a reserved random access slot, which has not been used due to busy traffic or area condition.

[0038] Fig. 12 shows an example for a decay based cancellation of random access slots. The perspective view in Fig. 12 thereby shows five sequential second sections 3 provided every 10ms. The case shown in Fig. 12 is a subcase to the net base station originated cancellation shown in Fig. 11. The reserved random access slots 24 are reduced over a number of succeeding random access time windows based upon a schedule known to both the mobile station and the base station. In the case shown in Fig. 12, the mobile station reserves three random access time slots 24 in the first two succeeding random access time windows. In the third random access time window, the mobile station only reserves two random access time slots 24 and in the fourth random access time window, the mobile station only reserves one random access time slot 24. Conse-

quently, in the fifth succeeding random access time window, the corresponding random access time slot 25 is not reserved any more and available to other mobile stations. This application may be particularly applicable to bursty data which may require a larger number of initial random access slots to improve the initial access, but slower over a time period. This may occur for terminal emulation software. User commands need a fast response, however, if these commands initiate a large data transfer, the speed of access will not be required.

[0039] Fig. 13 shows a block diagram of a base station 26 and two mobile stations 27, 28, incorporating the present invention. The base station 26 comprises an antenna 31, the first mobile station 27 comprises an antenna 29 and the second mobile station 28 comprises an antenna 30. The base station 26 and the two mobile stations 27, 28 are arranged and adapted to transmit and receive data in a code division multiple access telecommunication system according to the present invention as described above relating to Figs. 2 - 12. The base station 26 and the two mobile stations 27, 28 thereby comprise the known elements, which are required for a proper operation. Thus, the base station 26 and the mobile stations 27, 28 according to the present invention comprise coders, encoders, interleavers, deinterleavers, modulators, demodulators, HF-receiving and transmission means and so on.

[0040] Further on, as shown in the block diagrams of Fig. 14 and 15, the base station 26 and the mobile stations comprise the corresponding means for incorporating the present invention. Thus, the mobile station 27, which is shown in Fig. 14, comprises an antenna 29 coupled to a receiving means 32 and a transmitting means 33. The receiving means 32 is coupled with a means 35 for determining the partition point transmitted with partitioning information from a base station. The receiving means 32 receives the partitioning information and the means 35 for determining the partition point extracts the partition point from the received partitioning information. The means 34 for choosing a random access slot within a first section 2 of a random access time window is coupled to said means 35 for determining the partition point and also to a transmitting means 33 coupled to the antenna 29. The mobile station therefore knows from the received partitioning information, in which way the random access time windows are divided into the first section 2 and the second section 3, and the means 34 for choosing the random access slot correspondingly chooses a random access slot from the first section 2 to transmit random access data to the corresponding mobile station. The receiving means 32, the transmitting means 33, the means 34 for choosing random access slots and the means 35 for determining the partition point are further connected to the known means required for normal operation of a mobile station.

[0041] In Fig. 15, a block diagram of the base station 26 incorporating the present invention is shown. The base station 26 comprises a receiving means 36 and a transmitting means 37, which are both coupled to the antenna 31. A means 38 for partitioning the random access time windows in a first and a second section 2, 3 is coupled to the transmitting means 37. The means 38 for partitioning the random access time windows is thereby provided with information on the corresponding system requirements, on the basis of which it determines the partition point. The partitioning information is then transmitted over the transmitting means 37 and the antenna 31 to the corresponding mobile station, for example the mobile station 27. The means 38 for partitioning the random access time windows 1 sets the number of random access slots in the first and the second section 2, 3 variably depending on the system requirements and the transmitting means 37 transmits the partitioning information periodically to the connected mobile stations. The receiving means 36 receives random access data in one of the random access slots from the first sections 2 of random access time windows transmitted by a mobile station. The receiving means 36 is coupled to a means 39 for reserving random access slots in the second section 3 of the random access time windows in case that a random access burst transmitted in a contention based random access slot contains reservation information. The means 39 for reserving reservation based random access slots thereby extracts the reservation information from the received random access burst. Further, the means 39 for reserving reservation based random access slots generates a confirmation signal which is transmitted over the transmitting means 37 to the respective mobile station to confirm the reservation or to inform the mobile station, that no or fewer random access slots as required are reserved. The means 39 for reserving random access slots is thereby connected to the means 38 for defining the partition point, so that each time a new partition point is defined, information on the currently reserved random access slots can be respected. The base station 26 further comprises a means 40 for determining a non-usage of reserved random access slots in case that the base station operates according to the example shown in Fig. 11. Upon the determination of a non-usage of a predetermined number of reserved random access slots, the means 40 generates a corresponding cancellation signal, which is transmitted over the transmission means 37 to the mobile station, as described relating to Fig. 11.

[0042] The advantage of the method and the devices according to the present invention is, that a big part of the contention based access is removed. Further, an efficient use of the random access channel and therefore the random access time windows is ensured, since, when contention based use is not in demand, more capacity can be given to reserved resources. Thereby, the partitioning can be changed dynamically by the network. Further, the random access bursts can carry additional information embodied as either user or control data. This can be used to describe the type of

channel required or to make a future reservation. Another advantage of the present invention is that changes to the layer 1 standards are not required. Further, the percentage partitioning of the random access time window can be employed optionally be a network provider. If he wishes not to use the capability, the number of reservation based random access slots can be set to zero.

[0043] Although in the above description the partitioning of a random access time window into two sections has been described, partitioning of a random access time window into three or more sections is possible. Further, as an additional feature to the partitioning of the random access time window, the random access slots within the different sections of the random access time window can be divided into groups having different priority classes. Since the random access data to be transmitted in the random access slots can have different transmission priorities, it can be ensured, that some types of random access data will have a better possibility of gaining access to the network due to the higher probability that a particular message will be successful in reaching the base station. The network can dynamically change the size of the groups depending on system requirements and broadcasts the corresponding information to the mobile station.

Claims

1. Method for transmitting and receiving data in a code division multiple access telecommunication system, comprising the steps of

   providing a random access time window (1) comprising a plurality of random access slots for transmitting random access data from at least one first communication device (27, 28) to a second communication device (26),
   partitioning the random access time window (1) in a first and a second section (2, 3), whereby the first section contains contention based random access slots and the second section (3) contains reservation based random access slots.

2. Method for transmitting and receiving data according to claim 1,
   characterized in,

   that in said partitioning step the number of random access slots in the first and second section (2, 3), respectively, is set variably depending on system requirements.

3. Method for transmitting and receiving data according to claim 1 or 2,
   characterized in,

   that each random access slot in said random access time window is defined by a time offset value and a preamble code.

4. Method for transmitting and receiving data according to claim 1, 2 or 3,
   characterized in,

   that said second communication device periodically transmits partitioning information defining the partitioning of said random access time window to at least one first communication device (27, 28).

5. Method for transmitting and receiving data according to one of the preceding claims,
   characterized in,

   that a first communication device (27, 28) for transmitting random access data to said second communication device (26), randomly chooses one of said random access slots from said first section (2) of said random access time window (1).

6. Method for transmitting and receiving data according to claim 5,
   characterized in,

   that said random access data transmitted in a random access slot of said first section (2) comprise reservation data for reserving at least one random access slot of said second section (3) in at least one succeeding random access time window (1).

7. Method for transmitting and receiving data according to claim 6,
   characterized in,

   that said reservation data comprise information on the time duration required for the total number of reserved random access slots to indicate the end of the reservation.

8. Method for transmitting and receiving data according to claim 7,
   characterized in,

   that said reservation data comprise information on a number of random access slots to be reserved in the first succeeding random access time window (1) and information on a continuous reduction of said number in the following random access time windows.

9. Method for transmitting and receiving data according to one of the claims 6 to 8,

characterized in,

that said first communication device (27, 28), for indicating the end of the reservation of the required random access slots, transmits reservation termination data in said first or second section (2) of a random access time window (1).

10. Method for transmitting and receiving data according to one of the claims 6 to 8,
characterized in,

that said second communication device (26), upon determination of a non-usage of reserved random access slots, terminates the reservation.

11. Method for transmitting and receiving data according to claim 10,
characterized in,

that said second communication device (26) transmits information on the termination of the reservation to the corresponding first communication device.

12. Device (27, 28) for transmitting and receiving data in a code division multiple access telecommunication system, in which a random access time window (1) comprising a plurality of random access slots for transmitting random access data is provided, the random access time window being partitioned in a first and a second section (2, 3), whereby the first section (2) contains contention based random access slots and the second section (3) contains reservation based random access slots, with means (34) for randomly choosing a random access slot from said first section, and means (33) for transmitting random access data in said chosen random access slot.

13. Device for transmitting and receiving data according to claim 12,
characterized in,

that each random access slot in said random access time window is defined by a time offset value and a preamble code.

14. Device for transmitting and receiving data according to claim 12 or 13,
characterized by

means (32) for receiving periodically transmitted partitioning information defining the partitioning of said random access time window (1).

15. Device for transmitting and receiving data according to one of the claims 12 to 14,
characterized in,

that said random access data transmitted in a random access slot of said first section (2) comprise reservation data for reserving at least one random access slot of said second section (3) in at least one succeeding random access time window (1).

16. Device for transmitting and receiving data according to one of the claims 12 to 14,
characterized in,

that said reservation data comprise information on the time duration required for the total number of reserved random access slots to indicate the end of the reservation.

17. Device for transmitting and receiving data according to claim 16,
characterized in,

that said reservation data comprise information on a number of random access slots to be reserved in the first succeeding random access time window and information on a continuous reduction of said number in the following random access time windows.

18. Device for transmitting and receiving data according to one of the claims 12 to 17,
characterized in,

that said means (33) for transmitting random access data, for indicating the end of the reservation of the required random access slots, transmits reservation termination data in said first or second section (2) of a random access time window (1).

19. Device (26) for transmitting and receiving data in a code division multiple access telecommunication system, in which a random access time window (1) comprising a plurality of random access slots for transmitting random access data is provided, with means (38) for partitioning the random access time window in a first and a second section (2, 3), whereby the first section (2) contains contention based random access slots and the second section (3) contains reservation based random access slots, and means (37) for transmitting partitioning information defining the partitioning of said random access time window (1).

20. Device for transmitting and receiving data according to claim 19,

characterized in,

that said means (38) for partitioning the random access time window (1) sets the number of random access slots in said first and second section, respectively, variably depending on system requirements, and
said means (37) for transmitting partitioning information periodically transmits said partitioning information.

21. Device for transmitting and receiving data according to claim 19 or 20,
characterized by

means (36) for receiving random access data in one of said random access slots from said first section (2) of said random access time window (1).

22. Device for transmitting and receiving data according to claim 21,
characterized in,

that said random access data transmitted in a random access slot of said first section (2) comprise reservation data for reserving at least one random access slot of said second section (3) in at least one succeeding random access time window, and that means (39) for reserving at least one random access slot of said second section (3) in at least one succeeding random access time window are provided.

23. Device for transmitting and receiving data according to claim 22,
characterized in,

that said reservation data comprise information on the time duration required for the total number of reserved random access slots to indicate the end of the reservation.

24. Device for transmitting and receiving data according to claim 22 or 23,
characterized in,

that said reservation data comprise information on a number of random access slots to be reserved in the first succeeding random access time window and information on a continuous reduction of said number in the following random access time windows.

25. Device for transmitting and receiving data according to claim 22, 23 or 24,
characterized by

means (40) for determining of a non-usage of reserved random access slots, whereby the reservation is terminated upon a positive result of said determination.

26. Device for transmitting and receiving data according to claim 25,
characterized by

means (37) for transmitting information on the termination of the reservation upon a positive result of said determination.

FIG 1

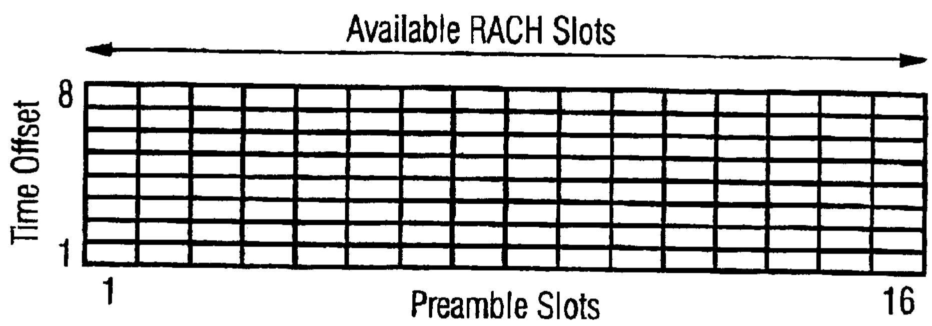


FIG 2

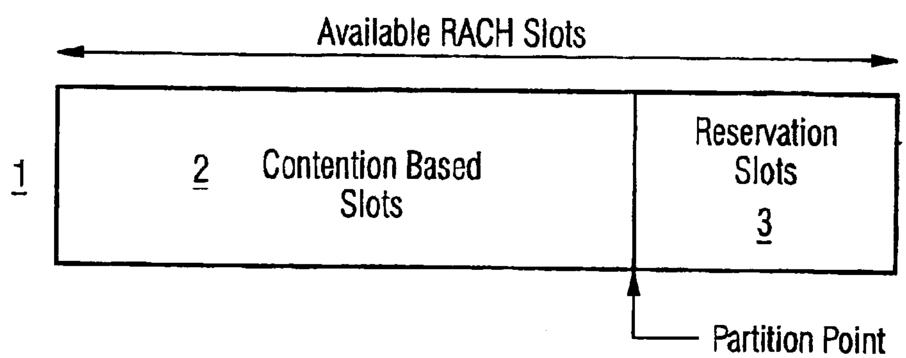


FIG 3

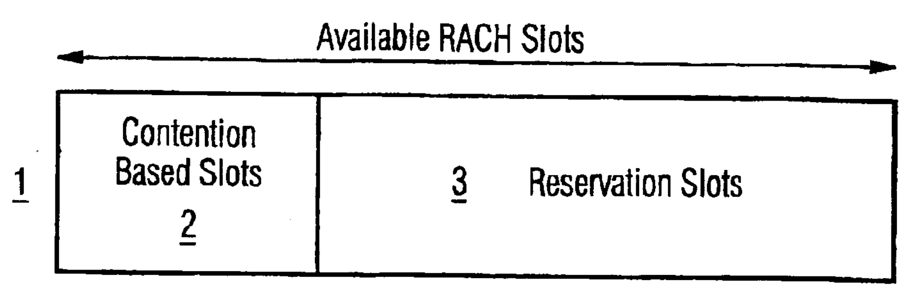


FIG 4

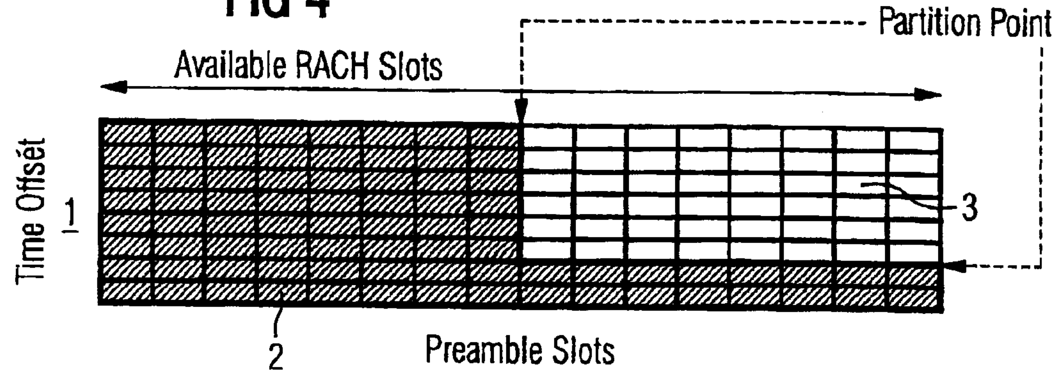


FIG 5

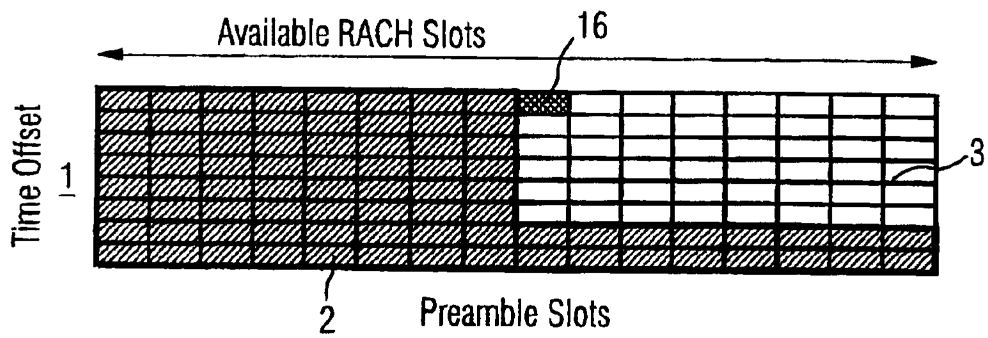


FIG 6

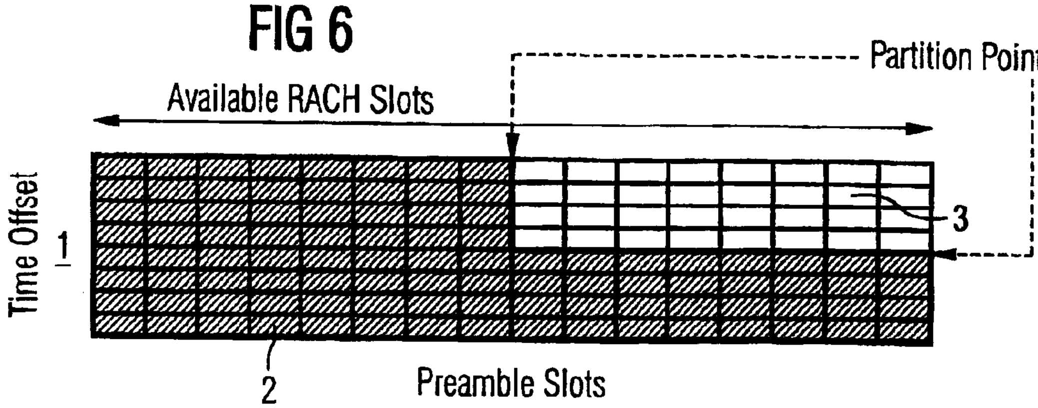


FIG 7

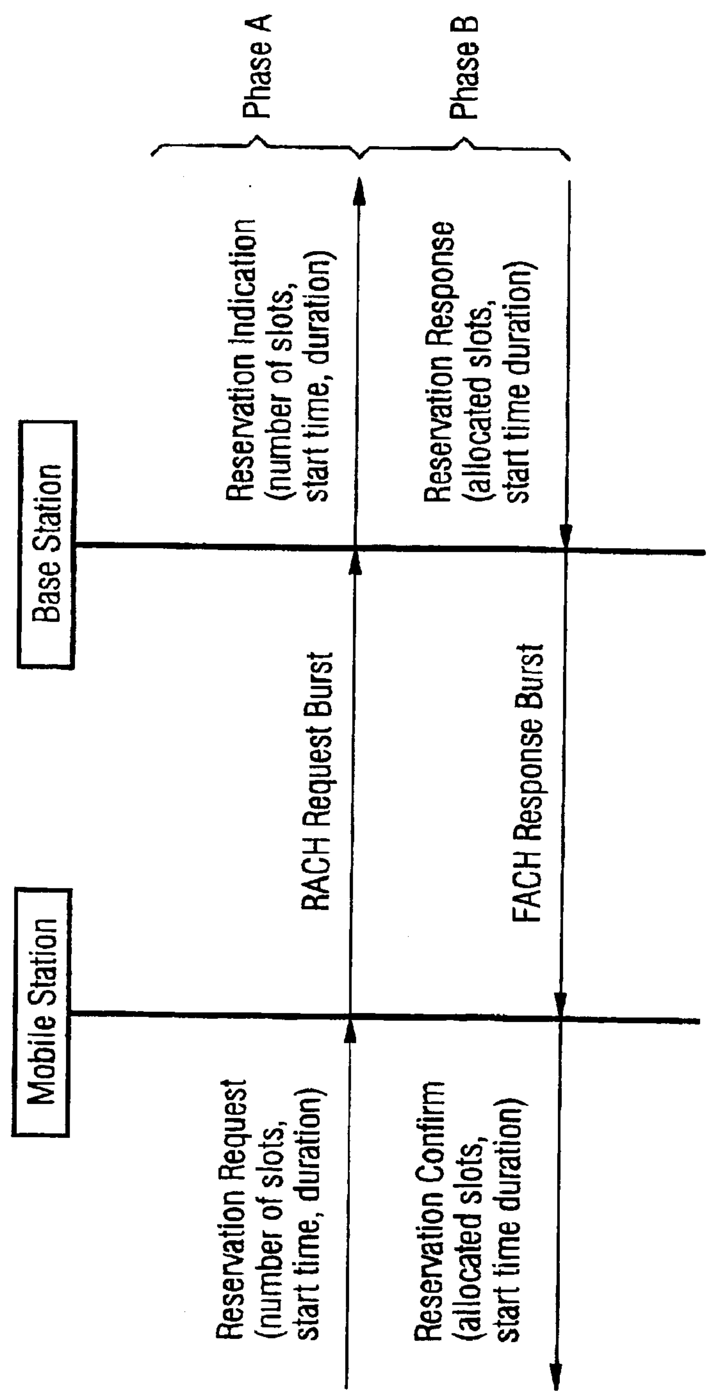

Mobile Station
Base Station
Reservation Request (number of slots, start time, duration)
RACH Request Burst
Reservation Indication (number of slots, start time, duration)
Phase A
Reservation Confirm (allocated slots, start time duration)
FACH Response Burst
Reservation Response (allocated slots, start time duration)
Phase B

FIG 8 A

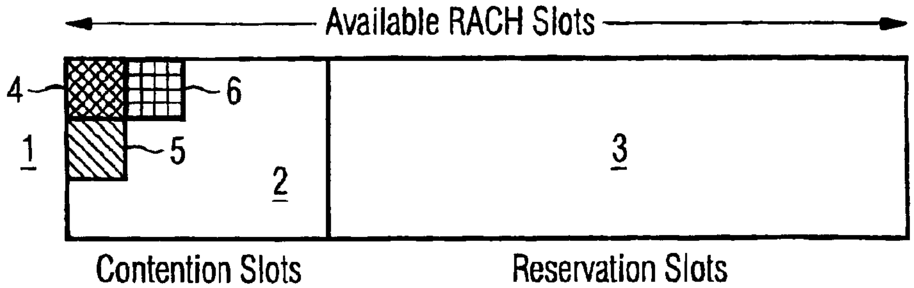


FIG 8 B

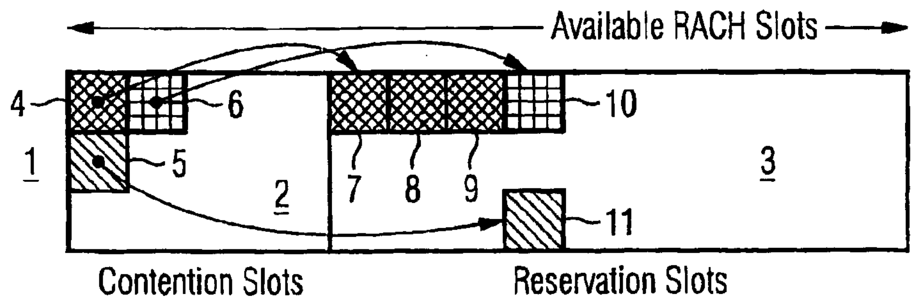


FIG 8 C

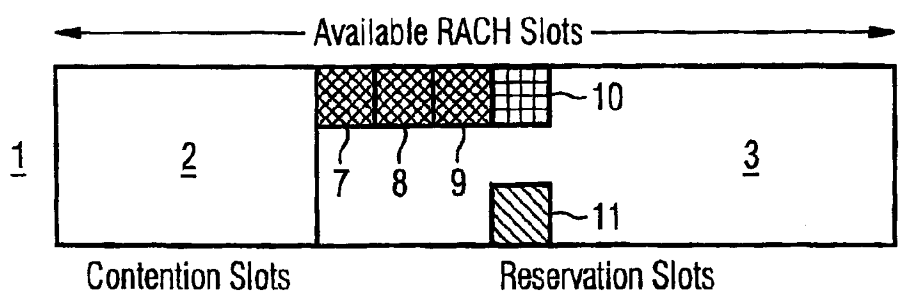


FIG 8 D

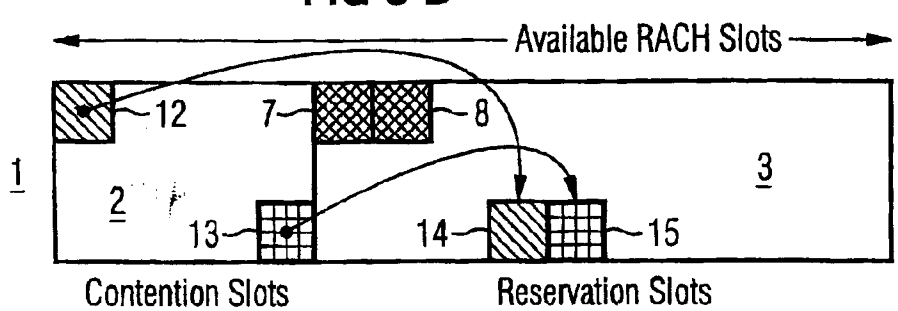


FIG 9

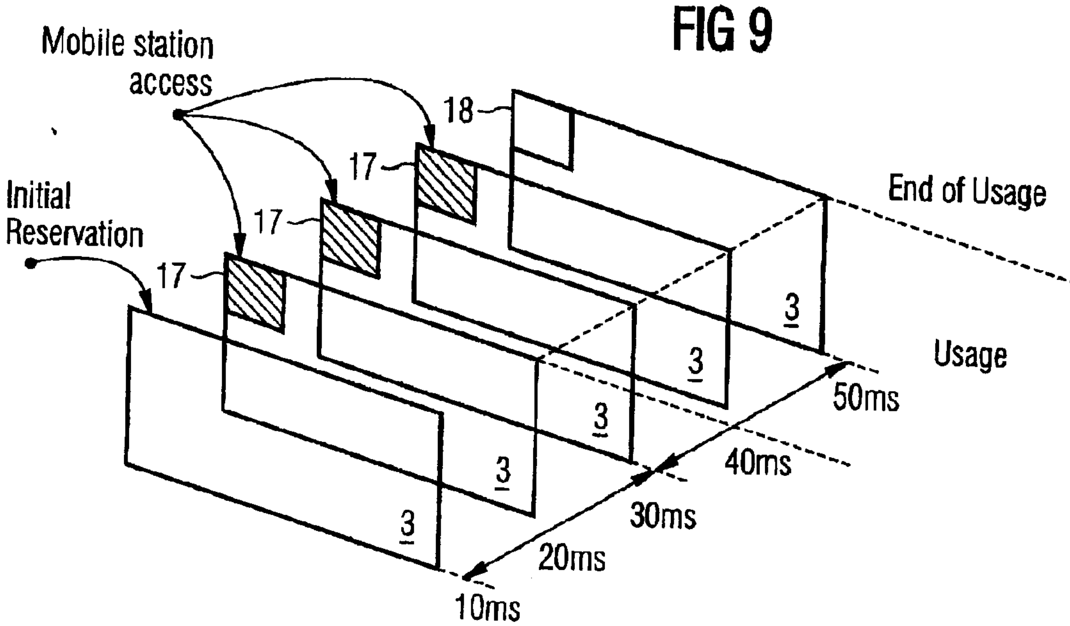

Mobile station access
Initial Reservation
18
17
17
17
End of Usage
Usage
3
3
3
3
3
50ms
40ms
30ms
20ms
10ms

FIG 10

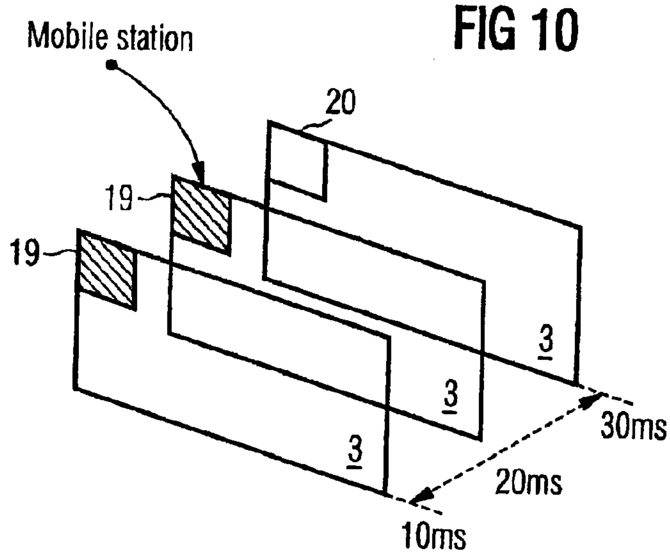

Mobile station
20
19
19
3
3
3
30ms
20ms
10ms

FIG 11

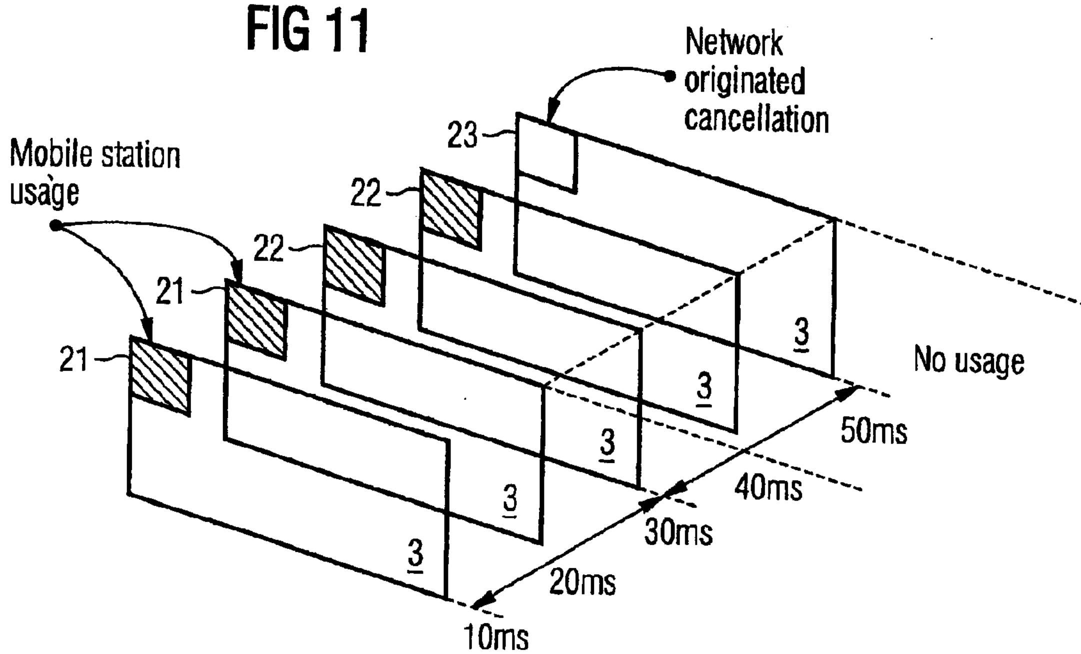

Network
originated
cancellation
Mobile station
usage
23
22
22
21
21
3
3
3
3
3
No usage
50ms
40ms
30ms
20ms
10ms

FIG 12

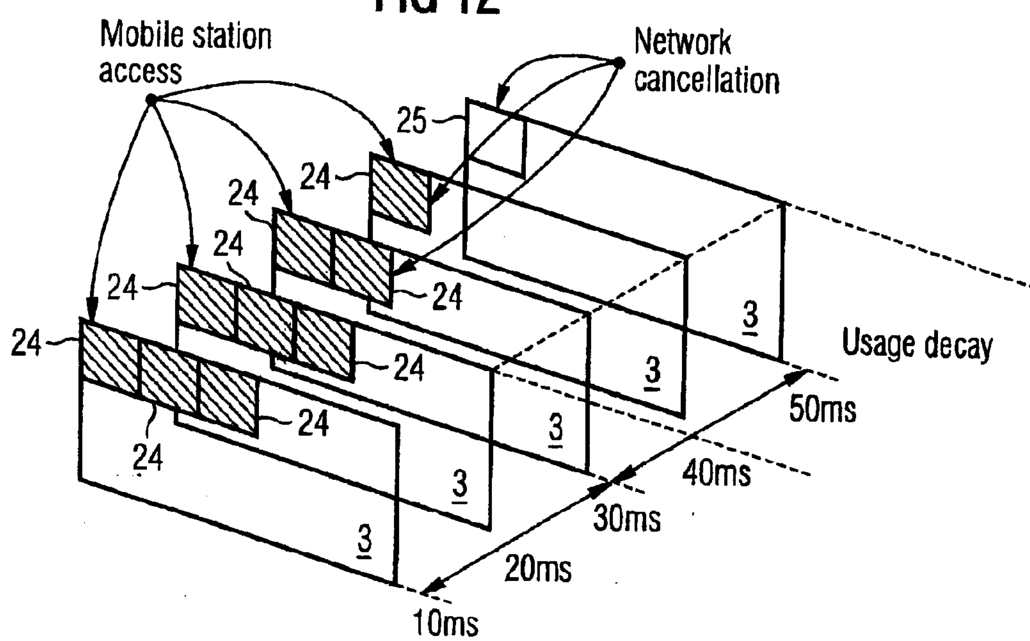

Mobile station
access
Network
cancellation
25
24
24
24
24
24
24
24
24
24
3
3
3
3
3
Usage decay
50ms
40ms
30ms
20ms
10ms

FIG 13

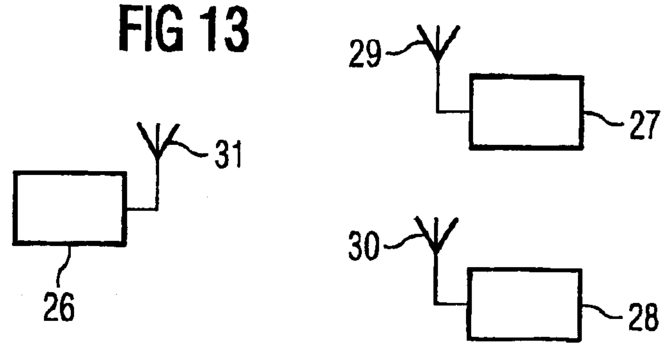


FIG 14

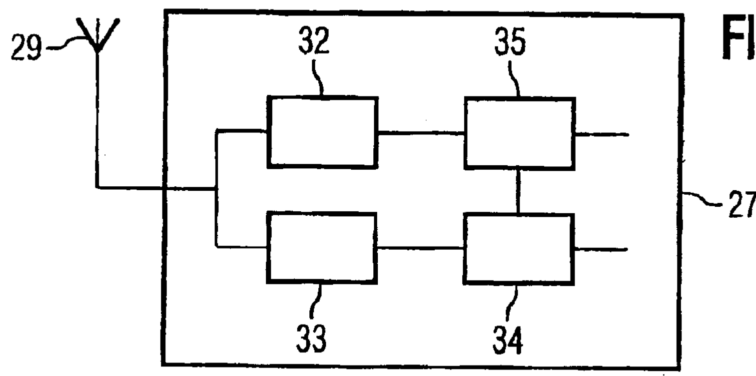
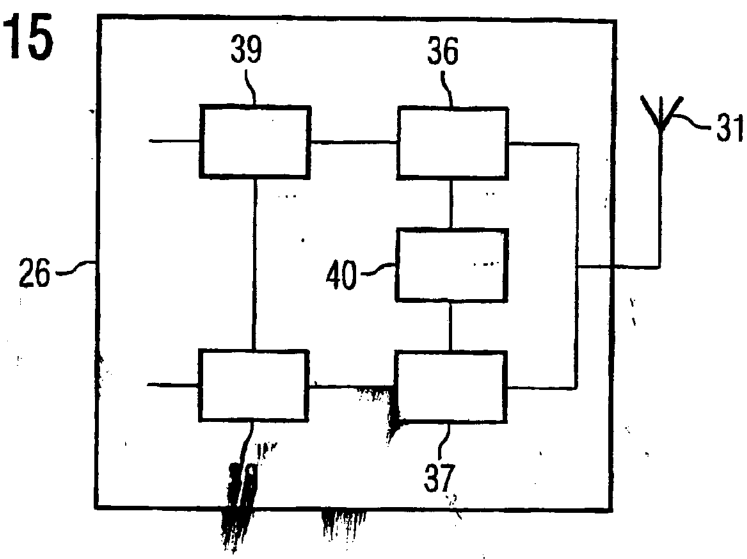


FIG 15



European Patent Office

EUROPEAN SEARCH REPORT

Application Number
EP 98 11 8816

DOCUMENTS CONSIDERED TO BE RELEVANT

Category	Citation of document with indication, where appropriate, of relevant passages	Relevant to claim	CLASSIFICATION OF THE APPLICATION (Int.Cl.6)
X	US 5 123 029 A (BANTZ DAVID F ET AL) 16 June 1992 * abstract; claims 1,3 * * column 2, line 66 - column 3, line 8 * * column 3, line 30 - line 35 * * column 5, line 52 - line 64 * * column 6, line 3 - line 9 * * column 6, line 28 - line 61; figures 3,3A * * column 8, line 17 - line 62 * ---	1,2,4,5, 12,14, 19-21	H04Q7/38 H04B7/216 H04L12/28
Y	EP 0 765 096 A (NIPPON TELEGRAPH & TELEPHONE) 26 March 1997 * abstract * * column 2, line 21 - line 40 * * column 3, line 31 - column 4, line 10 * * column 14, line 42 - column 15, line 16; figure 6 * ---	1,2,4-6, 12,14, 15,19-22	
Y	EP 0 621 708 A (IBM) 26 October 1994 * abstract; figure 3 * * page 3, line 56 - line 57 * * page 4, line 8 - line 11 * * page 4, line 52 - page 5, line 3 * * page 6, line 1 - line 56 * * page 7, line 55 - page 8, line 15 * * page 10, line 9 - line 26 * * page 11, line 28 - line 33 * ---	1,2,4-6, 12,14, 15,19-22	TECHNICAL FIELDS SEARCHED (Int.Cl.6) H04Q H04B H04L
A	WO 97 06622 A (PHILIPS ELECTRONICS NV ;PHILIPS NORDEN AB (SE); PHILIPS ELECTRONIQ) 20 February 1997 * abstract * * page 3, line 25 - page 5, line 19 * -----	1,2,12, 15,19,20	

The present search report has been drawn up for all claims

Place of search	Date of completion of the search	Examiner
THE HAGUE	25 March 1999	Harris, E

CATEGORY OF CITED DOCUMENTS

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25-03-1999

Patent document cited in search report	Publication date	Patent family member(s)	Publication date
US 5123029 A	16-06-1992	EP 0523362 A	20-01-1993
		JP 2101728 C	22-10-1996
		JP 5191345 A	30-07-1993
		JP 8013155 B	07-02-1996
EP 0765096 A	26-03-1997	JP 9093646 A	04-04-1997
		CN 1159110 A	10-09-1997
		JP 9233051 A	05-09-1997
EP 0621708 A	26-10-1994	US 5384777 A	24-01-1995
		BR 9401518 A	27-12-1994
		CA 2115211 A,C	20-10-1994
		CN 1100857 A	29-03-1995
		JP 2662181 B	08-10-1997
		JP 7015433 A	17-01-1995
		KR 138001 B	01-07-1998
WO 9706622 A	20-02-1997	CN 1165602 A	19-11-1997
		EP 0786181 A	30-07-1997
		JP 10507611 T	21-07-1998

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